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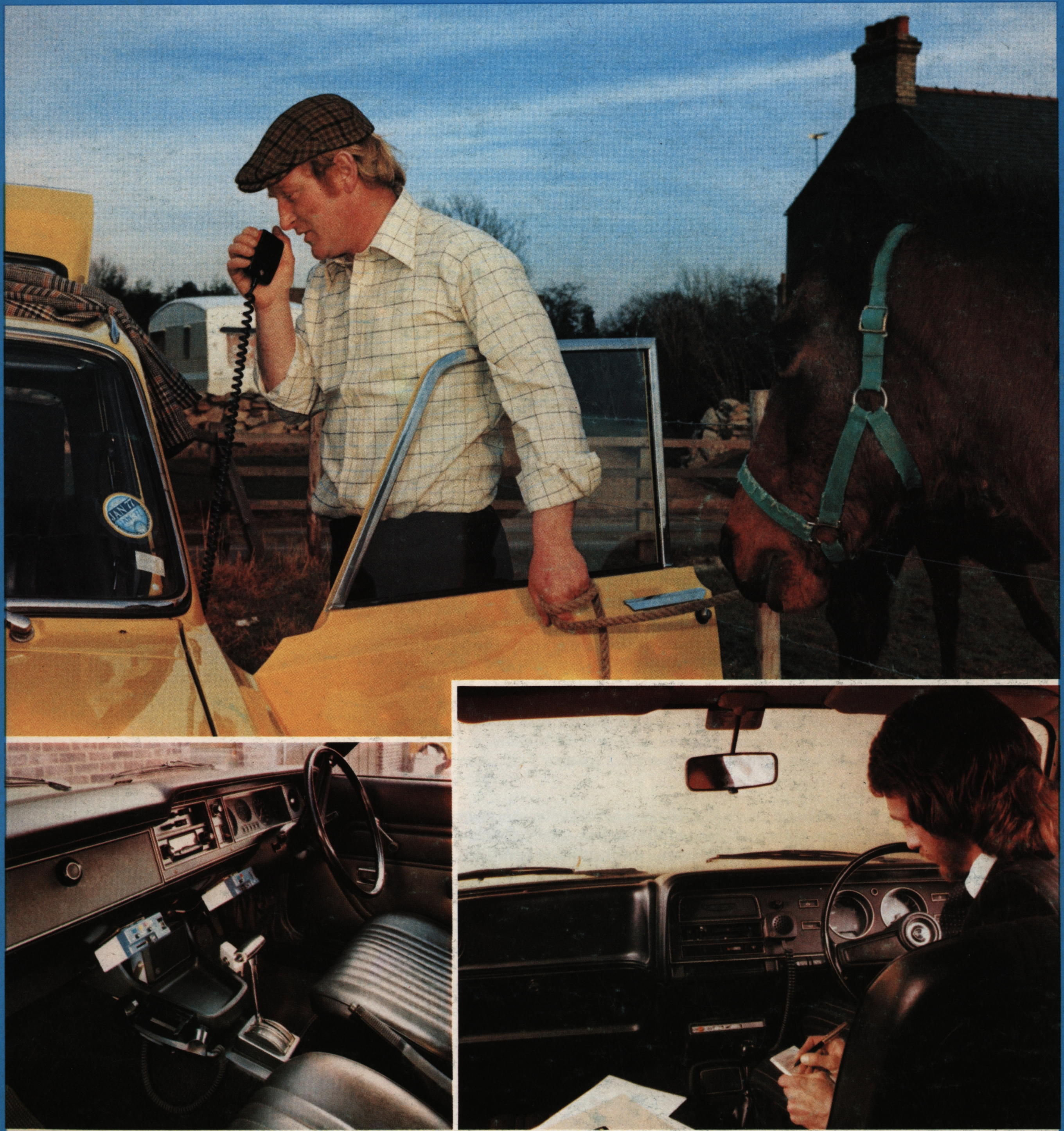

- The TDA1071 radio IC in communications receivers
- Zero-crossing detector circuits
- The TDA1050 IC in car radio a.m. receivers
- Simple temperature sensing circuits using thermistors
- Zinc oxide VDRs for transient suppression

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Cover picture:
Mobile communications are becoming increasingly important in many aspects of everyday life. This issue includes an article on the TDA1071, an integrated circuit designed specifically for mobile radio applications. (Cover photographs by courtesy of Pye Telecommunications Ltd).

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This journal records developments in the application of electronics in science, industry, and the home. It contains information on the properties and performance of solid-state devices, valves and tubes, and associated components and assemblies. The practical circuits given are designed to provide the best working conditions and performance for the devices used. The design principles may be used (so far as device ratings allow) for adaptation to meet particular requirements. Current data should be consulted before any circuit is considered for production. Mullard Limited can normally supply only those devices or assemblies listed in the Company's current catalogues. Availability in quantity should be confirmed.

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The TDA1071 radio IC in communications receivers

J.A. GARTERS and J.S. MALCOLM
Mullard Central Application Laboratory

This article describes a new radio IC, the TDA1071, which has been recently developed for use in a.m. or f.m. equipment. Four applications of the IC in communications receivers are described in detail, and examples of further applications briefly mentioned.

INTRODUCTION

The TDA1071 (development number M715) is a unique integrated circuit for use in mains and portable radio equipment. The low current consumption and supply voltage, together with a very low peripheral component count, make the IC particularly suitable for portable battery-operated equipment.

The integrated circuit incorporates an a.m. oscillator, an a.m. mixer with a.g.c., a four-stage differential amplifier and limiter, and a four-quadrant multiplier. The novelty of the IC lies in the combination of both f.m. and a.m. functions in the multiplier circuit, which acts as a quadrature detector (with squelch) on f.m., and as a synchronous demodulator on a.m.

A description of the IC is given, together with details of the application of the device in four configurations commonly found in communications receivers; the overall design of such receivers is simplified by the high sensitivity achieved using this IC. Further applications in Citizens Band and single-sideband equipment are also briefly described.

IC CIRCUIT DESCRIPTION

The complete internal circuit of the TDA1071 is shown in Fig.1; the three main sections are described below.

A.M. oscillator and mixer

The circuit of the a.m. oscillator and mixer (TR$_1$ to TR$_{13}$) is substantially the same as that used in the TBA570, details of which are given in Ref.1. The positive supply for this circuit (pin 11) is separate from the main supply to allow decoupling between the oscillator and the mains supply on a.m., and to allow the oscillator/mixer circuit to be switched off for f.m. operation.

Amplifier/limiter

The amplifier/limiter section consists of four differential amplifier stages, the outputs from each stage being buffered by emitter followers; the output of the second stage is brought out to pin 4 via a separate emitter follower (TR$_{38}$). The collector resistors of the stages are scaled to give a good limiting characteristic, and stray coupling between stages is minimised by providing a separate buffered supply to each stage (via TR$_{47}$ to TR$_{50}$). The tail currents for the differential amplifiers and the emitter currents for the emitter followers are supplied by current sources TR$_{14}$ to TR$_{29}$, forming a current mirror with TR$_{81}$ and TR$_{82}$. The d.c. conditions are stabilised by overall d.c. feedback via R$_{58}$.

Multiplier

Transistors TR$_{56}$ to TR$_{59}$, TR$_{65}$, and TR$_{66}$ form a conventional four-quadrant multiplier circuit, supplied by current source TR$_{68}$ acting as a current mirror with TR$_{67}$. The bases of TR$_{56}$ to TR$_{59}$ are fed by emitter followers TR$_{55}$ and TR$_{60}$, the bases of which are brought out to pins 8 and 9. The bases of TR$_{65}$ and TR$_{66}$ are connected to the outputs of the differential amplifier/limiter circuit, and the emitter resistors of these transistors are included to improve circuit balance.

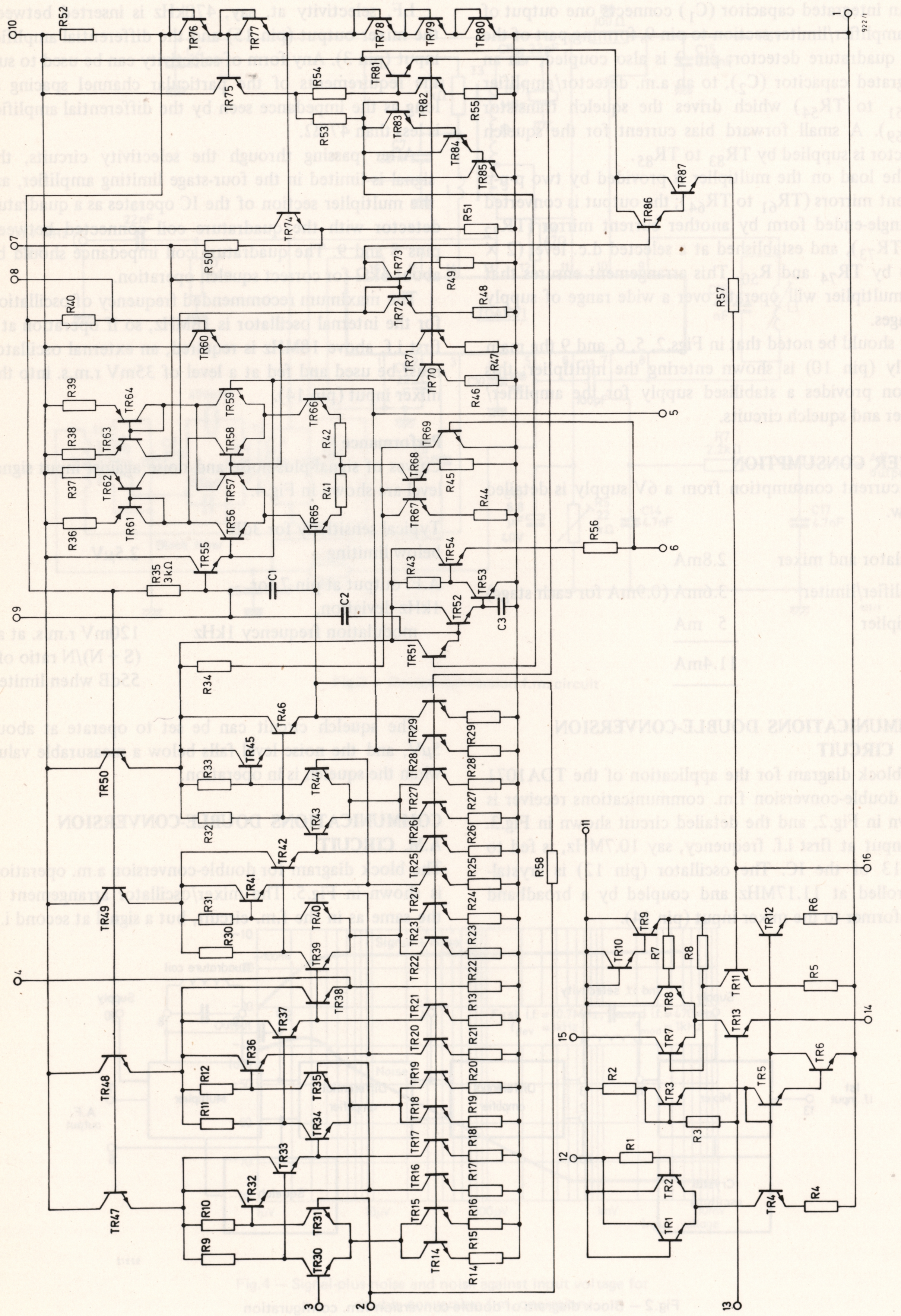

Fig.1 – Internal circuit of the TDA1071

An integrated capacitor (C_1) connects one output of the amplifier/limiter section to pin 9, forming part of the f.m. quadrature detector; pin 9 is also coupled, via an integrated capacitor (C_2), to an a.m. detector/amplifier (TR_{51} to TR_{54}) which drives the squelch transistor (TR_{69}). A small forward bias current for the squelch detector is supplied by TR_{83} to TR_{85}.

The load on the multiplier is provided by two p-n-p current mirrors (TR_{61} to TR_{64}); the output is converted to single-ended form by another current mirror (TR_{72} and TR_{73}), and established at a selected d.c. level ($2 \times V_{be}$) by TR_{74} and R_{50}. This arrangement ensures that the multiplier will operate over a wide range of supply voltages.

It should be noted that in Figs.2, 5, 6, and 9 the main supply (pin 10) is shown entering the multiplier; this section provides a stabilised supply for the amplifier/limiter and squelch circuits.

POWER CONSUMPTION

The current consumption from a 6V supply is detailed below.

Oscillator and mixer	2.8mA
Amplifier/limiter	3.6mA (0.9mA for each stage)
Multiplier	5 mA
	11.4mA

COMMUNICATIONS DOUBLE-CONVERSION F.M. CIRCUIT

The block diagram for the application of the TDA1071 in a double-conversion f.m. communications receiver is shown in Fig.2, and the detailed circuit shown in Fig.3. An input at first i.f. frequency, say 10.7MHz, is fed to pin 13 of the IC. The oscillator (pin 12) is crystal-controlled at 11.17MHz and coupled by a broadband transformer to the mixer input (pin 14).

I.F. selectivity at, say, 470kHz is inserted between the mixer output (pin 15) and the differential amplifier input (pin 3). Any form of selectivity can be used to suit the requirements of the channel spacing as long as the impedance seen by the differential amplifier is less than 470Ω.

After passing through the selectivity circuits, the signal is limited in the four-stage limiting amplifier, and the multiplier section of the IC operates as a quadrature detector with the quadrature coil connected between pins 8 and 9. The quadrature coil impedance should be about 6kΩ for correct squelch operation.

The maximum recommended frequency of oscillation for the internal oscillator is 18MHz, so if operation at a first i.f. above 18MHz is required, an external oscillator must be used and fed at a level of 35mV r.m.s. into the mixer input (pin 14).

Performance

Curves of signal-plus-noise and noise against input signal level are shown in Fig.4.

Typical sensitivity for 3dB below limiting	2.5µV
A.F. output at pin 7 for 1kHz deviation, modulation frequency 1kHz	120mV r.m.s. at a (S + N)/N ratio of 55dB when limited

The squelch circuit can be set to operate at about 5µV, and the noise level falls below a measurable value when the squelch is in operation.

COMMUNICATIONS DOUBLE-CONVERSION A.M. CIRCUIT

The block diagram for double-conversion a.m. operation is shown in Fig.5. The mixer/oscillator arrangement is the same as in the f.m. circuit, but a signal at second i.f.

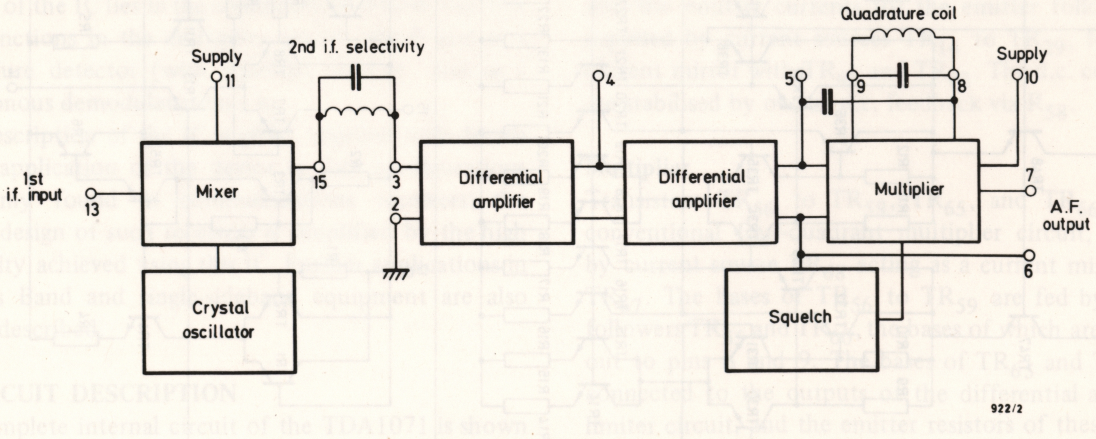
Quadrature coil
Supply
11
2nd i.f. selectivity
4
5
9
8
Supply
10
1st i.f. input
13
Mixer
15
3
2
Differential amplifier
Differential amplifier
Multiplier
7
A.F. output
6
Crystal oscillator
Squelch
922/2

Fig.2 — Block diagram of double-conversion f.m. configuration

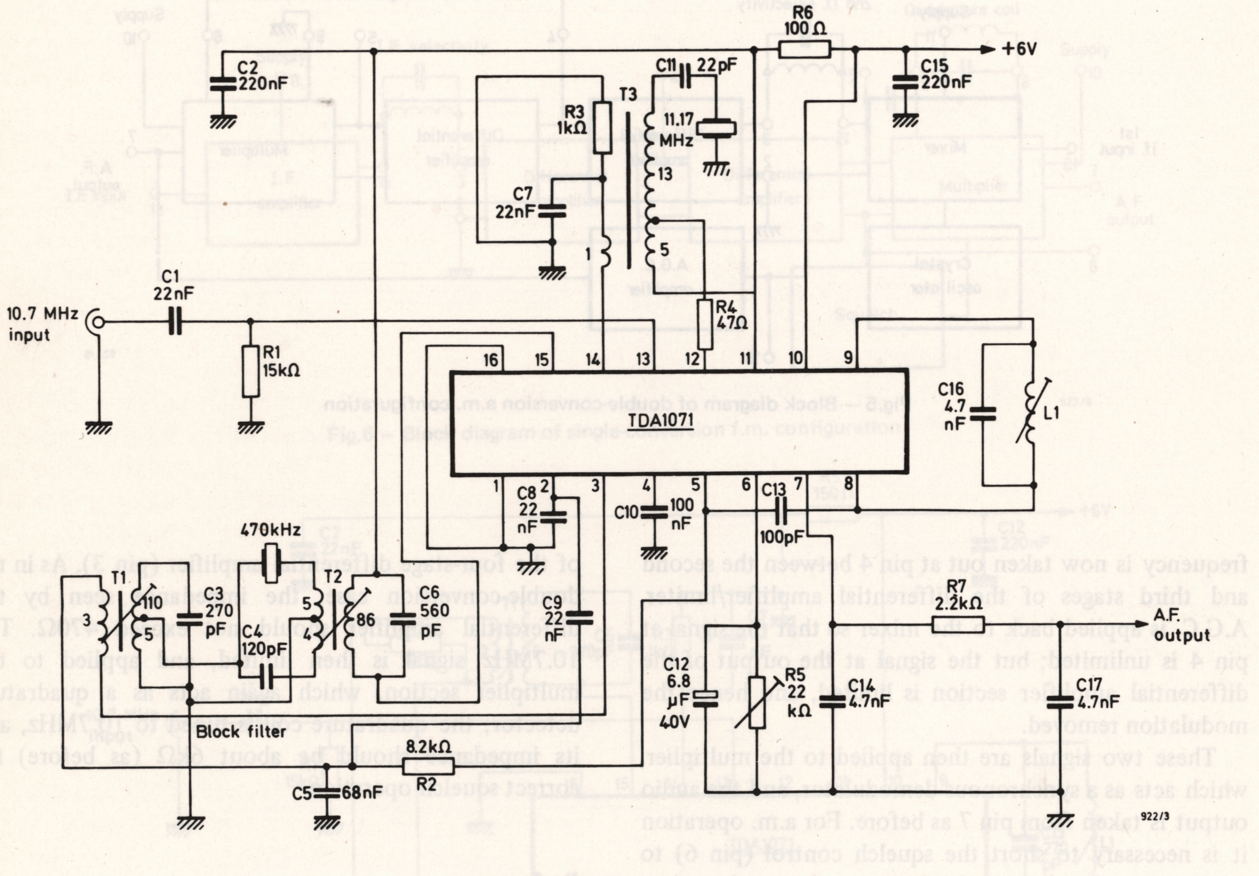
R6
100Ω
+6V
C2
220nF
C11 22pF
C15
220nF
R3
1kΩ
T3
11.17 MHz
13
C7
22nF
1
5
C1
22nF
10.7 MHz input
R1
15kΩ
R4
47Ω
16 15 14 13 12 11 10 9
TDA1071
1 2 3 4 5 6 7 8
C16
4.7 nF
L1
C8
22 nF
C10
100 nF
C13
100pF
470kHz
T1
3
110
5
C3
270 pF
C4
120pF
T2
5
4
86
C6
560 pF
C9
22 nF
Block filter
R7
2.2kΩ
A.F. output
C12
6.8 µF
40V
R5
22 kΩ
C14
4.7nF
C17
4.7nF
8.2kΩ
R2
C5 68nF
922/3

Fig.3 — Double-conversion f.m. circuit

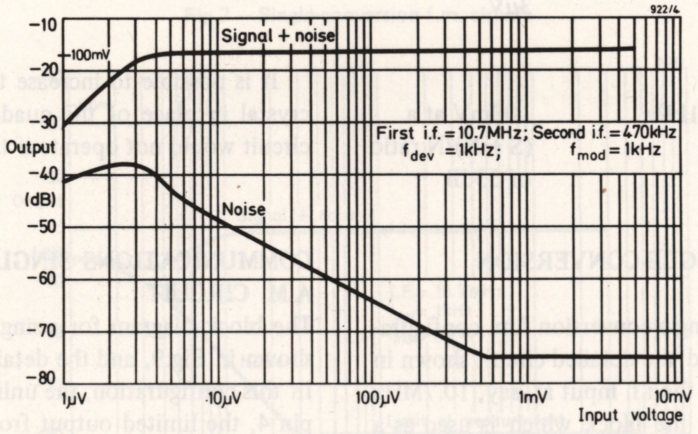

Fig.4 — Signal-plus-noise and noise against input voltage for double-conversion f.m. operation

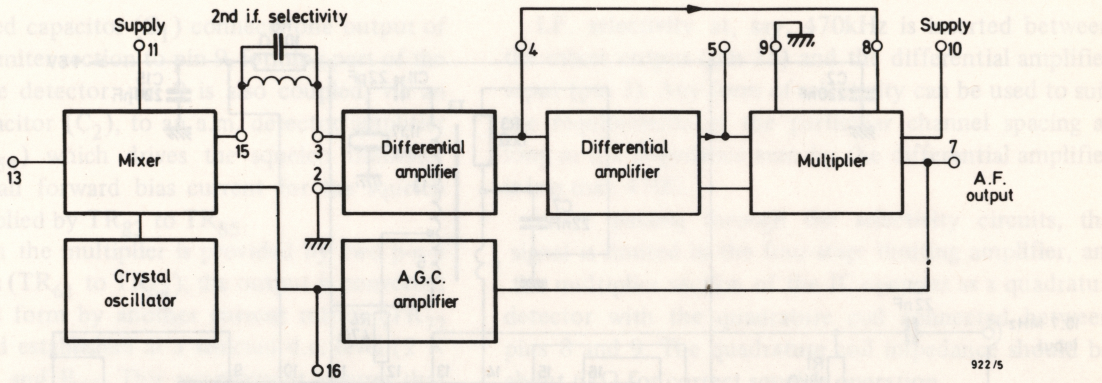
Supply
11
2nd i.f. selectivity
1st i.f. input
13
Mixer
15
3
2
Differential amplifier
Crystal oscillator
A.G.C. amplifier
16
4
5
9
8
Supply
10
Differential amplifier
Multiplier
7
A.F. output
922/5

Fig.5 — Block diagram of double-conversion a.m. configuration

frequency is now taken out at pin 4 between the second and third stages of the differential amplifier/limiter. A.G.C. is applied back to the mixer so that the signal at pin 4 is unlimited; but the signal at the output of the differential amplifier section is limited, and hence the modulation removed.

These two signals are then applied to the multiplier, which acts as a synchronous demodulator, and the audio output is taken from pin 7 as before. For a.m. operation it is necessary to short the squelch control (pin 6) to earth; a.g.c. for front-end gain control may be taken from pin 16.

As this configuration is normally of less interest than the f.m. system, no separate circuit diagram or performance curve is given. However, the circuit of Fig.3 may be converted to a.m. operation by making the modifications described above.

Performance

Typical sensitivity for a (S + N)/N ratio of 10dB	3μV
A.F. output at pin 7 for 30% modulation, modulation frequency 1kHz	100mV at a (S + N)/N ratio of 53dB

COMMUNICATIONS SINGLE-CONVERSION F.M. CIRCUIT

The block diagram for a single-conversion f.m. configuration is shown in Fig.6, and the detailed circuit shown in Fig.7. In this f.m. system, the i.f. input at, say, 10.7MHz is applied to pin 13, and the block which is used as a mixer in the double-conversion systems already described is now used as an extra i.f. amplifier with an output at pin 15. This output is fed via a tuned circuit to the input of the four-stage differential amplifier (pin 3). As in the double-conversion case, the impedance seen by the differential amplifier should not exceed 470Ω. The 10.7MHz signal is then limited, and applied to the multiplier section, which again acts as a quadrature detector; the quadrature coil is tuned to 10.7MHz, and its impedance should be about 6kΩ (as before) for correct squelch operation.

Performance

Curves of signal-plus-noise and noise against input signal level for the single-conversion f.m. configuration are shown in Fig.8.

Typical sensitivity for 3dB below limiting	5μV
A.F. output at pin 7 for 1kHz deviation, modulation frequency 1kHz	6mV r.m.s. at a (S + N)/N ratio of 35dB when limited

It is possible to increase the audio output by using a crystal in place of the quadrature coil, but the squelch circuit would not operate if this were done.

COMMUNICATIONS SINGLE-CONVERSION A.M. CIRCUIT

The block diagram for a single-conversion a.m. system is shown in Fig.9, and the detailed circuit shown in Fig.10. In this configuration, the unlimited output is taken from pin 4, the limited output from pin 5, and the multiplier acts as a synchronous demodulator. A.G.C. is applied to the input stage i.f. amplifier, which is the section used as the mixer in the double-conversion systems.

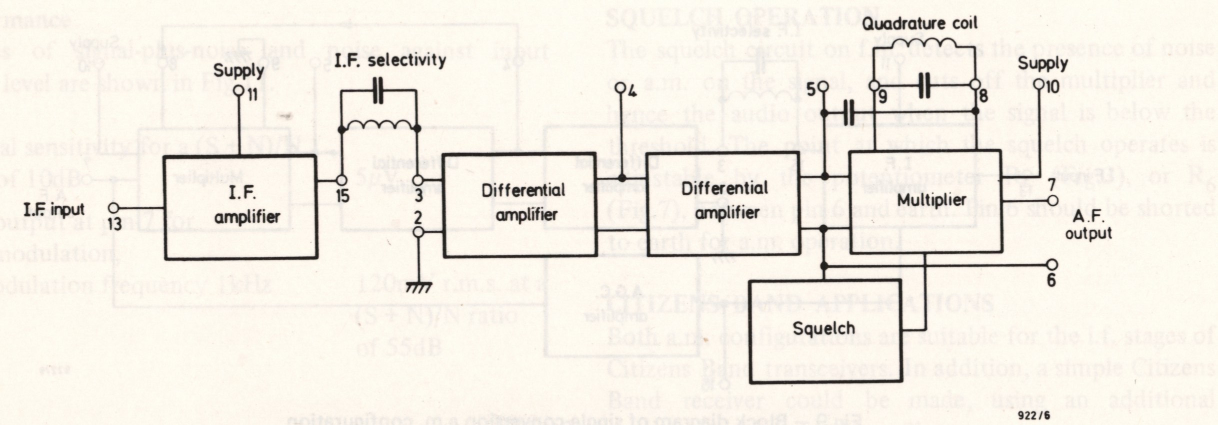
Supply
11
I.F. selectivity
I.F input
13
I.F. amplifier
15
3
2
Differential amplifier
4
5
9
Quadrature coil
8
Supply
10
Differential amplifier
Multiplier
7
A.F. output
6
Squelch
922/6

Fig.6 — Block diagram of single-conversion f.m. configuration

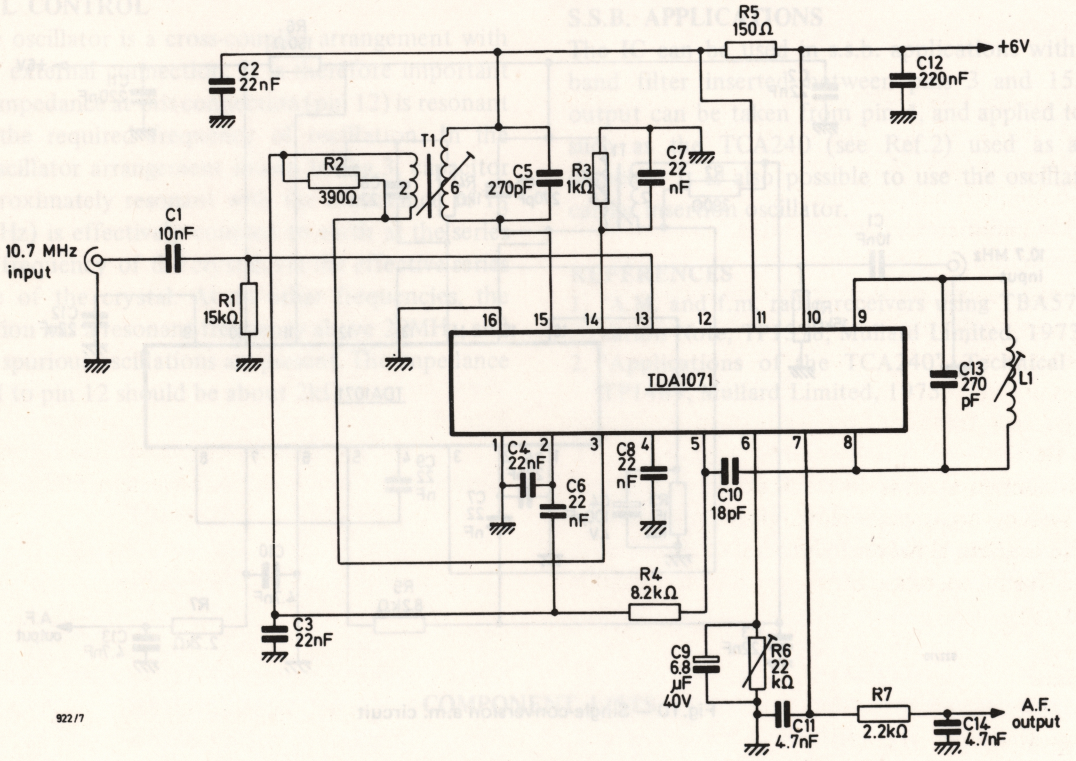
R5
150Ω
+6V
C12
220nF
C2
22nF
R2
390Ω
T1
1
2
3
6
C5
270pF
R3
1kΩ
C7
22
nF
C1
10nF
10.7 MHz
input
R1
15kΩ
16
15
14
13
12
11
10
9
TDA1071
C13
270
pF
L1
1
2
3
4
5
6
7
8
C4
22nF
C6
22
nF
C8
22
nF
C10
18pF
R4
8.2kΩ
C3
22nF
C9
6.8
μF
40V
R6
22
kΩ
C11
4.7nF
R7
2.2kΩ
C14
4.7nF
A.F.
output
922/7

Fig.7 — Single-conversion f.m. circuit

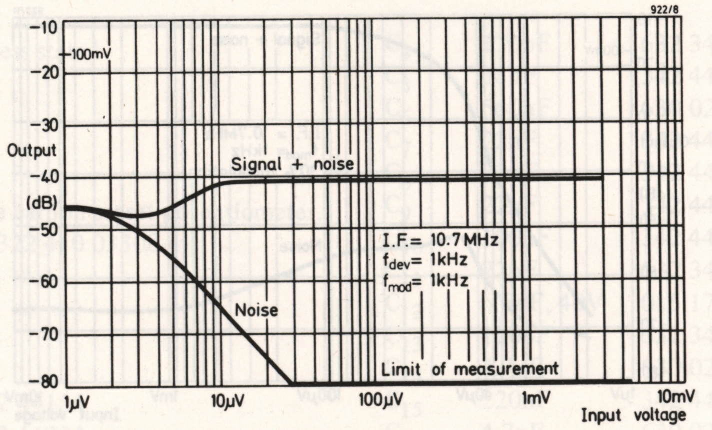

Fig.8 — Signal-plus-noise and noise against input voltage for single-conversion f.m. operation

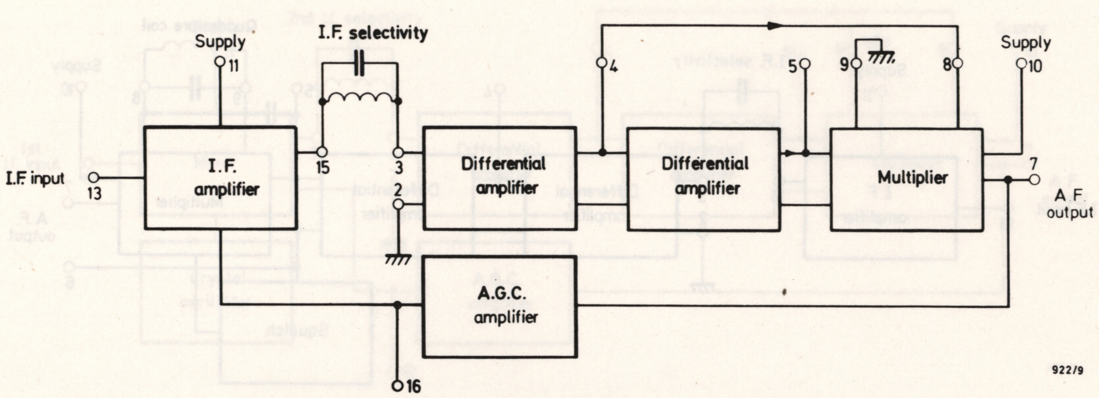
Supply
11
I.F. selectivity
4
5
9
8
Supply
10
I.F. input
13
I.F. amplifier
15
3
2
Differential amplifier
Differential amplifier
Multiplier
7
A.F. output
A.G.C. amplifier
16
922/9

Fig.9 — Block diagram of single-conversion a.m. configuration

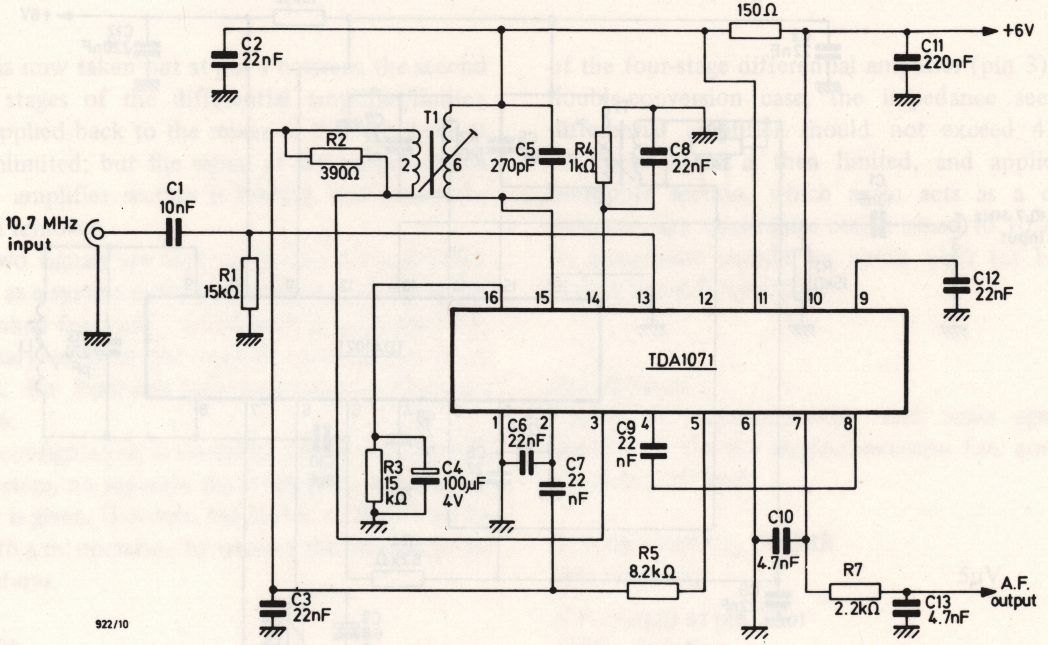
R6
150Ω
+6V
C2
22nF
C11
220nF
T1
R2
390Ω
2
6
C5
270pF
R4
1kΩ
C8
22nF
C1
10nF
10.7 MHz
input
R1
15kΩ
C12
22nF
16
15
14
13
12
11
10
9
TDA1071
1
C6
22nF
2
3
C9
22
nF
4
5
6
7
8
R3
15
kΩ
C4
100µF
4V
C7
22
nF
C10
4.7nF
R5
8.2kΩ
R7
2.2kΩ
A.F.
output
C3
22nF
C13
4.7nF
922/10

Fig.10 — Single-conversion a.m. circuit

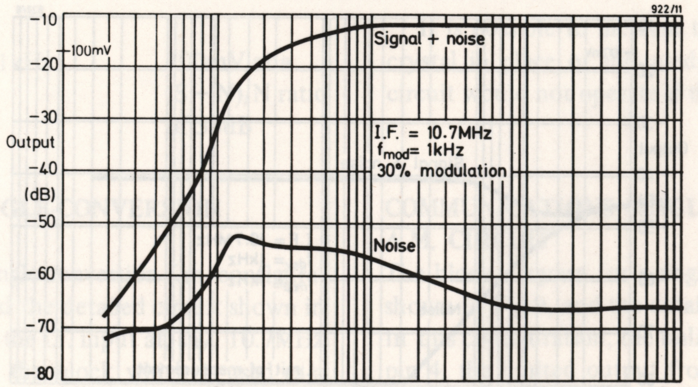

Fig.11 — Signal-plus-noise and noise against input voltage for single-conversion a.m. operation

Performance

Curves of signal-plus-noise and noise against input signal level are shown in Fig.11.

Typical sensitivity for a (S + N)/N ratio of 10dB	5µV
A.F. output at pin 7 for 30% modulation, modulation frequency 1kHz	120mV r.m.s. at a (S + N)/N ratio of 55dB

USE OF THE INTERNAL OSCILLATOR WITH CRYSTAL CONTROL

The basic oscillator is a cross-coupled arrangement with only one external connection. It is therefore important that the impedance at this connection (pin 12) is resonant only at the required frequency of oscillation. In the crystal oscillator arrangement shown in Fig.3, capacitor C_{11} (approximately resonant with the inductance of T_3 at 10.7MHz) is effectively coupled to earth at the series resonant frequency of the crystal, via the effective series resistance of the crystal. At all other frequencies, the combination has a resonant frequency above 20MHz, and hence no spurious oscillations are present. The impedance presented to pin 12 should be about 2kΩ.

SQUELCH OPERATION

The squelch circuit on f.m. detects the presence of noise or a.m. on the signal, and cuts off the multiplier and hence the audio output when the signal is below the threshold. The point at which the squelch operates is adjustable by the potentiometer R_5 (Fig.3), or R_6 (Fig.7), between pin 6 and earth. Pin 6 should be shorted to earth for a.m. operation.

CITIZENS BAND APPLICATIONS

Both a.m. configurations are suitable for the i.f. stages of Citizens Band transceivers. In addition, a simple Citizens Band receiver could be made, using an additional transistor as an external oscillator.

S.S.B. APPLICATIONS

The IC can be used in s.s.b. applications with the sideband filter inserted between pins 3 and 15. The i.f. output can be taken from pin 4, and applied to a device such as the TCA240 (see Ref.2) used as a product detector. It is also possible to use the oscillator as the carrier insertion oscillator.

REFERENCES

1. 'A.M. and f.m. radio receivers using TBA570', Application Note, TP1378, Mullard Limited, 1973.
2. 'Applications of the TCA240', Technical Note 18, TP1489, Mullard Limited, 1975.

COMPONENT LISTS

Double-conversion f.m. circuit (Fig.3)

Resistors

All resistors CR25 10% unless stated

R_1	15kΩ	
R_2	8.2kΩ	
R_3	1kΩ	
R_4	47Ω	
R_5	22kΩ	Miniature carbon preset potentiometer, Philips 2322 410 03308
R_6	100Ω	
R_7	2.2kΩ	

Capacitors

C_1	22nF	342 44154
C_2	220nF	342 44224
C_3	270pF	632 58221
C_4	120pF	632 34121
C_5	68nF	342 44683
C_6	560pF	630 02561
C_7	22nF	342 44154
C_8	22nF	342 44154
C_9	22nF	342 44154
C_{10}	100nF	342 44104
C_{11}	22pF	632 34229
C_{12}	6.8µF, 40V	015 17688
C_{13}	100pF	632 34101
C_{14}	4.7nF	630 02472
C_{15}	220nF	342 44224
C_{16}	4.7nF	630 02472
C_{17}	4.7nF	630 02472

Winding data

T_1	Primary: 115 turns, 0.071mm enamelled copper, tapped at 5 turns Secondary: 3 turns, 0.071mm enamelled copper, Toko 7P former 7XN(C)
T_2	Primary: 86 turns, 0.071mm enamelled copper Secondary: 9 turns, 0.071mm enamelled copper, tapped at 4 turns, Toko 7P former 7XN(C)
T_3	Primary: 18 turns, 0.315mm enamelled copper, tapped at 5 turns Secondary: 1 turn, 0.315mm enamelled copper, FX3850 toroid (see Fig.12)
L_1	30 turns, 0.071mm enamelled copper, Toko 7P former 7XN(C)

Single-conversion f.m. circuit (Fig.7)

Resistors

All resistors CR25 10% unless stated

R_1	15kΩ	
R_2	390Ω	
R_3	1kΩ	
R_4	8.2kΩ	
R_5	150Ω	
R_6	22kΩ	Miniature carbon preset potentiometer, Philips 2322 410 03308
R_7	2.2kΩ	

Capacitors

C_1	10nF	342 44103
C_2	22nF	342 44154
C_3	22nF	342 44154
C_4	22nF	342 44154
C_5	270pF	632 58271
C_6	22nF	342 44154
C_7	22nF	342 44154
C_8	22nF	342 44154
C_9	6.8μF, 40V	015 17688
C_{10}	18pF	632 10189
C_{11}	4.7nF	630 02472
C_{12}	220nF	342 44224
C_{13}	270pF	632 58271
C_{14}	4.7nF	630 02472

Winding data

T_1	Primary: 6 turns, 0.16mm enamelled copper Secondary: 2 turns, 0.16mm enamelled copper, Toko 7P former 119AN(C)
L_1	6 turns, 0.16mm enamelled copper, Toko 7P former 119AN(C)

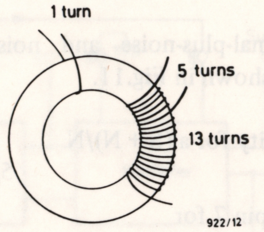

Fig.12 — Winding data of transformer T_3 in Fig.3 (see Component list)

Single-conversion a.m. circuit (Fig.10)

Resistors

All resistors CR25 10%

R_1	15kΩ
R_2	390Ω
R_3	15kΩ
R_4	1kΩ
R_5	8.2kΩ
R_6	150Ω
R_7	2.2kΩ

Capacitors

C_1	10nF	342 44103
C_2	22nF	342 44154
C_3	22nF	342 44154
C_4	100μF, 4V	015 12101
C_5	270pF	632 58271
C_6	22nF	342 44154
C_7	22nF	342 44154
C_8	22nF	342 44154
C_9	22nF	342 44154
C_{10}	4.7nF	630 02472
C_{11}	220nF	342 44224
C_{12}	22nF	342 44154
C_{13}	4.7nF	630 02472

Winding data

T_1	Primary: 6 turns, 0.16mm enamelled copper Secondary: 2 turns, 0.16mm enamelled copper, Toko 7P former 119AN(C)

Zero-crossing detector circuits

D.R. ARMSTRONG
Mullard Central Application Laboratory

This article describes a number of simple circuits for the detection of zero-voltage reference points on the mains supply waveform, and discusses their suitability for various power control applications. The circuits illustrate stages in the development of a fairly sophisticated circuit which may be used in phase control applications where the accuracy of zero-crossing detection is critical. In each case, appropriate component values are given and the performance of the circuit illustrated.

INTRODUCTION

Many power control circuits operating from the mains supply include a solid-state switch, such as a triac or thyristor, as the controlling element. Such a switch requires a trigger pulse at a single time, or at several times, in each mains cycle, the timing of the pulse being related to the mains waveform.

In the case of a.c. phase control, the trigger pulse should be symmetrically placed in the positive and negative half-cycles of the mains supply, and this requires a means of defining some datum point in each half-cycle. The most easily detected common point on the waveform is the zero-crossing, or zero-voltage point; a circuit to detect this point is therefore an important aspect of power control in washing machines, electric drills, lamp dimmers, and other applications requiring phase control. Symmetry of the trigger pulse is less important in applications requiring burst firing of the power supply, such as heating appliances, and so the zero-crossing detector circuit need not be as complex as that required for phase control. An example of the use of a simple zero-crossing detector in a control circuit is given in Ref.1.

Various circuits are described, the choice of a particular circuit depending on the symmetry and duration of pulse required for a particular application.

CIRCUIT DETAILS

In each of the circuits described below, it is assumed that a 12V supply is available; most circuits which require a zero-crossing detector will include a suitable stabilised low-voltage supply. All component values given are nominal, and all calculations are based on these nominal values.

Basic circuit

The basic zero-crossing detector circuit is shown in Fig.1. Transistors TR_1 and TR_2 are connected in common-emitter and common-base configurations respectively; TR_1 conducts when the mains voltage is positive, and TR_2 conducts when it is negative. An output pulse is

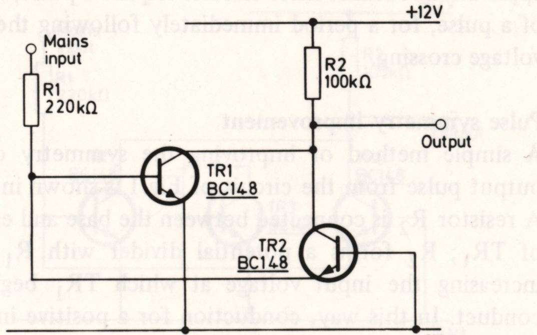

Fig.1 — Basic zero-crossing detector circuit

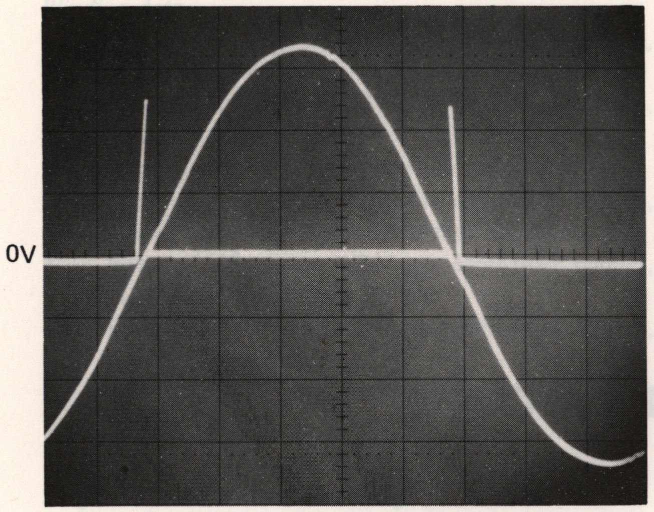

Fig.2 — Input and output waveforms of the circuit of Fig.1
Mains input 100V per division
Output pulse 5V per division
Timebase 2ms per division

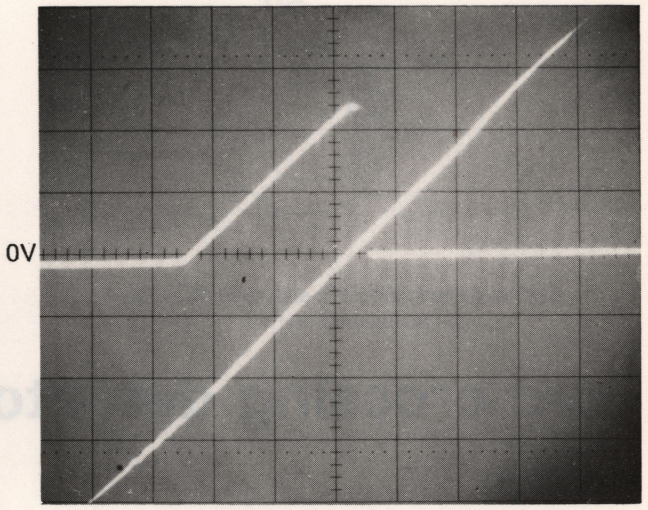

Fig.3 — Detail of input and output waveforms of Fig.2
Mains input 10V per division
Output pulse 5V per division
Timebase 100µs per division

therefore produced when the input waveform passes through zero, as shown in Fig.2 which gives the output waveform superimposed on the input waveform.

However, if the output waveform is examined more closely, it can be seen in Figs.3 and 4 that the pulse is not generated symmetrically about the zero-voltage point of the input. This is because of the different configurations of the transistors, the gain of the circuit being 125 to 900 for a positive input and only about unity for a negative input; hence the conduction will be greater when the mains supply is positive, and the pulse will be generated almost entirely when the input is negative.

In phase control applications, an asymmetrical pulse will lead to a d.c. component appearing in the load. This component may be acceptable in such applications as lamp dimmers, but will cause considerable problems in circuits with inductive loads or transformer inputs.

Asymmetry also leads to problems in burst-firing applications because these circuits require a pulse, or part of a pulse, for a period immediately following the zero-voltage crossing.

Pulse symmetry improvement

A simple method of improving the symmetry of the output pulse from the circuit of Fig.1 is shown in Fig.5. A resistor R_2 is connected between the base and emitter of TR_1; R_2 forms a potential divider with R_1, thus increasing the input voltage at which TR_1 begins to conduct. In this way, conduction for a positive input is delayed and the effective gain of the circuit for a positive input is reduced.

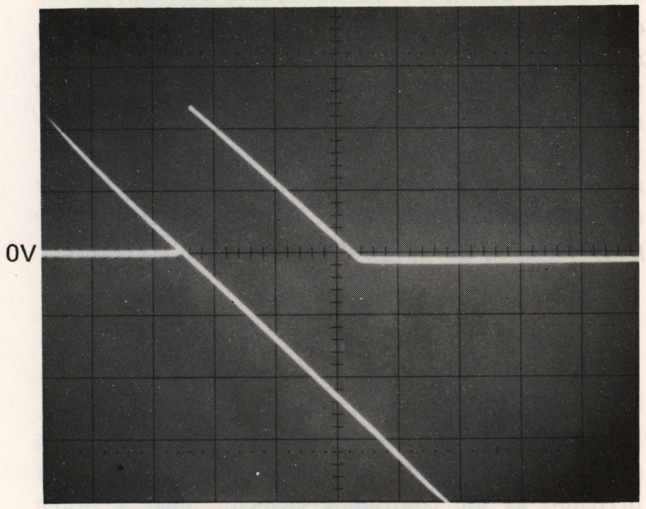

Fig.4 — Detail of input and output waveforms of Fig.2
Mains input 10V per division
Output pulse 5V per division
Timebase 100µs per division

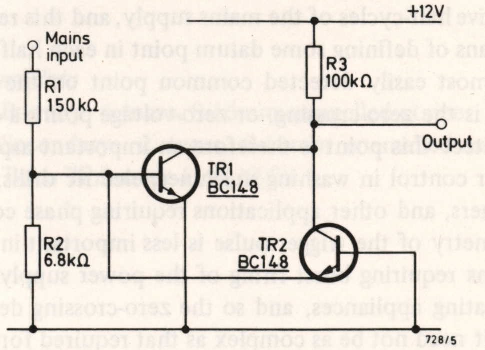

Fig.5 — Circuit to give improved pulse symmetry

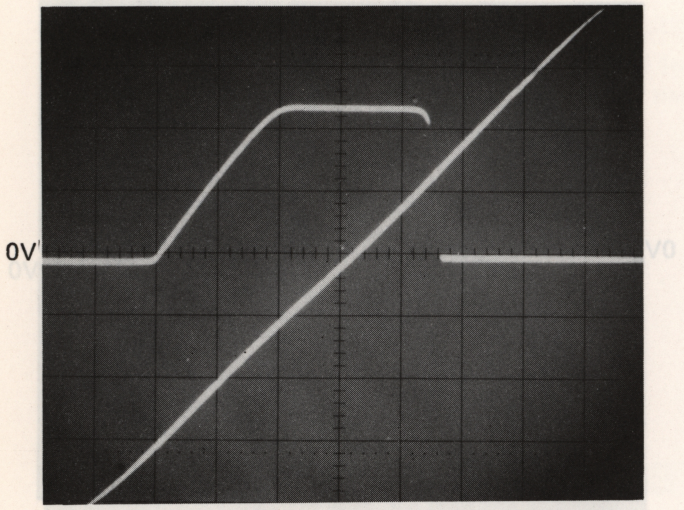

Fig.6 — Detail of input and output waveforms of the circuit of Fig.5.
Mains input 10V per division
Output pulse 5V per division
Timebase 100µs per division

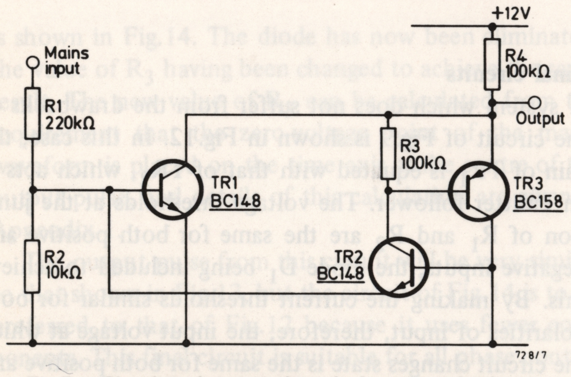

Fig.7 — Circuit to increase the gain for a negative input

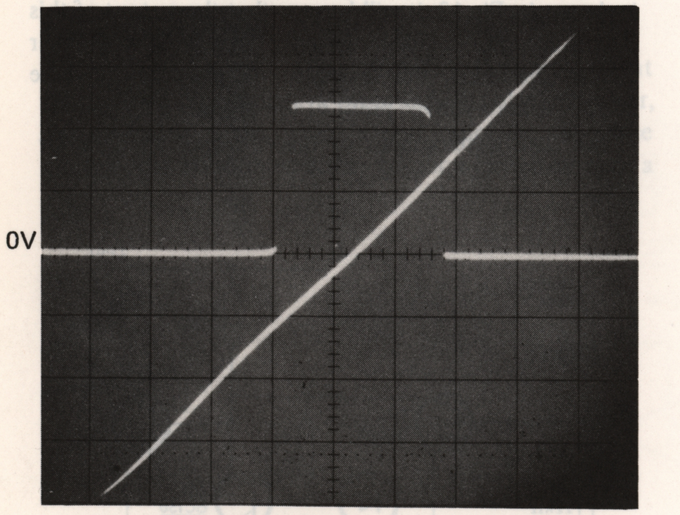

Fig.8 — Detail of input and output waveforms of the circuit of Fig.7.
Mains input 10V per division
Output pulse 5V per division
Timebase 100µs per division

A detail of the superimposed output and input waveforms is shown in Fig.6; only one detail is shown because the pulse at the negative-going zero-voltage point is similar to a mirror-image of the one shown. It can be seen that the delay in conduction for a positive input leads to a more symmetrical pulse, some portion of the pulse appearing when the input voltage is positive.

This improvement in symmetry will be sufficient for burst-firing control circuits, but may not be adequate for a.c. phase control applications where pulse symmetry is more important.

Further improvement is obtained by using the arrangement shown in Fig.7. The potential divider formed by R_1 and R_2 acts as before to increase the threshold voltages, but in addition a transistor, TR_3, is added to increase the sensitivity of the circuit to a negative input. In this circuit, the composite transistor formed by TR_2 and TR_3 has much the same gain as TR_1, a great improvement over the basic circuit which had a very asymmetrical transfer characteristic. A detail of the output and input waveforms for this circuit is shown in Fig.8. In this case, the pulse is almost perfectly symmetrical about the zero-crossing point, the error in the example being about 10µs.

Current mirror circuit

As an alternative to the circuit of Fig.7, in which an additional transistor was added to increase the gain of TR_2, the circuit shown in Fig.9 may be used. In this arrangement, transistor TR_3 is connected so as to reduce the gain of TR_1; these two transistors form a current mirror, which is shown in detail in Fig.10. The current mirror depends for its operation on the emitter-base voltages being the same for similar transistors conducting at the same current and at the same temperature; from this it is possible to calculate the current gain of the circuit of Fig.10. Thus:

$$I_{out} + I_B = I_E, \qquad \ldots (1)$$

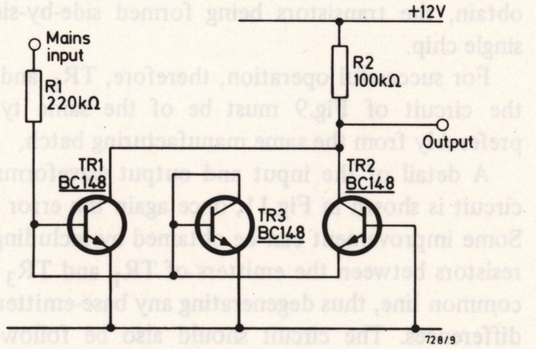

Fig.9 — Improved circuit using a current mirror

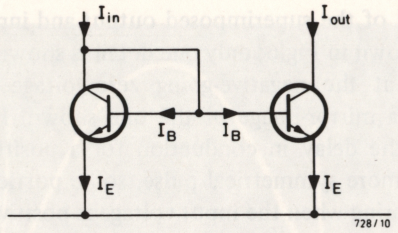

Fig.10 — Current mirror arrangement

and:

$$I_{in} - I_B = I_E. \qquad \ldots(2)$$

Assuming I_E is the same for both transistors, it may be eliminated between Eqs.1 and 2. Therefore:

$$I_{out} = I_{in} - 2I_B, \qquad \ldots(3)$$

but I_{out} is the collector current of one transistor; therefore:

$$I_{out} = h_{FE}I_B, \qquad \ldots(4)$$

where h_{FE} is the current gain of each transistor.

Substituting from Eq.4 into Eq.3 gives:

$$I_{out} = I_{in} - 2(I_{out}/h_{FE}), \qquad \ldots(5)$$

or:

$$\frac{I_{out}}{I_{in}} = \frac{1}{1 + 2/h_{FE}} = \frac{h_{FE}}{h_{FE} + 2}. \qquad \ldots(6)$$

From Eq.6 it can be seen that the gain of the circuit is reduced by a factor $1/(h_{FE} + 2)$ by the addition of a second transistor connected as shown.

Because the current mirror circuit requires two transistors with very similar base-emitter characteristics, it is not normally found in discrete circuits. However, it is a common feature in many integrated circuits, where good matching of characteristics is relatively easy to obtain, the transistors being formed side-by-side on a single chip.

For successful operation, therefore, TR_1 and TR_3 in the circuit of Fig.9 must be of the same type, and preferably from the same manufacturing batch.

A detail of the input and output waveforms of this circuit is shown in Fig.11; once again the error is small. Some improvement can be obtained by including similar resistors between the emitters of TR_1 and TR_3 and the common line, thus degenerating any base-emitter voltage differences. The circuit should also be followed by a squaring circuit to produce a better pulse shape.

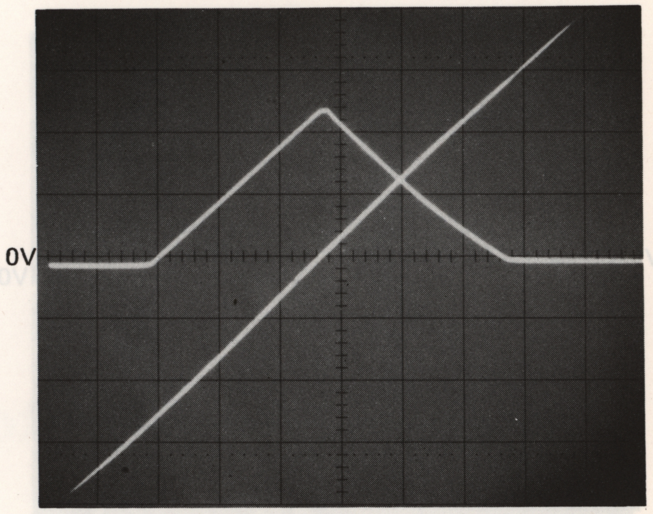

Fig.11 — Detail of input and output waveforms of the circuit of Fig.9.

Mains input	10V per division
Output pulse	5V per division
Timebase	100μs per division

Final circuits

A system which does not suffer from the drawbacks of the circuit of Fig.9 is shown in Fig.12. In this case, the gain of TR_1 is equated with that of TR_2, which acts as an emitter follower. The voltage thresholds at the junction of R_1 and R_2 are the same for both positive and negative inputs, the diode D_1 being included to achieve this. By making the current thresholds similar for both polarities of input, therefore, the input voltage at which the circuit changes state is the same for both positive and negative inputs.

A detail of the resultant input and output waveforms is shown in Fig.13; it will be seen that the output of this circuit has the opposite polarity to that of the other circuits discussed, and that the symmetry of the pulse about the zero-voltage point of the input is very good.

The final development of the zero-crossing detector is shown in Fig.14. The diode has now been eliminated, the value of R_3 having been changed to achieve the same result. The new value of R_3 can be calculated from the requirement that the zero-voltage point of the mains waveform is placed on the time axis at the centre of the output pulse, and details of this calculation are given in Appendix 1.

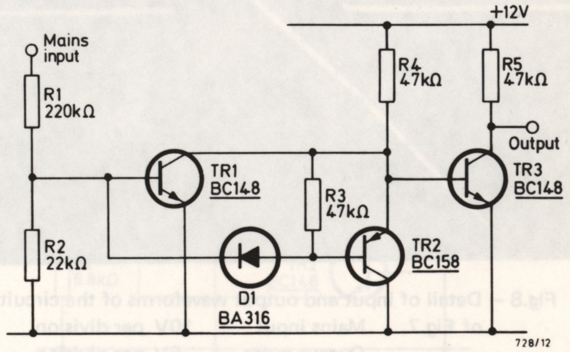

Fig.12 — Improved circuit to avoid transistor matching difficulties

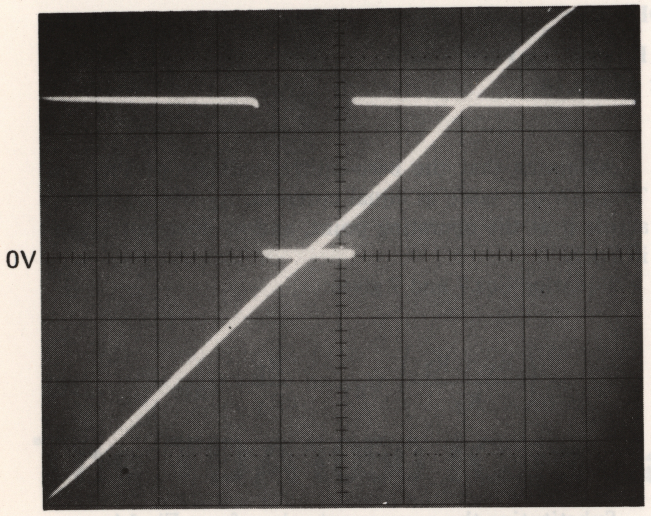

Fig.13 — Detail of input and output waveforms of the circuit of Fig.12.

Mains input	10V per division
Output pulse	5V per division
Timebase	100μs per division

The output pulse from this circuit will be very similar to that shown in Fig.13, but the circuit of Fig.14 is to be preferred to that of Fig.12 because it uses fewer components. This final circuit is suitable for all phase control applications and for burst-firing applications where a well-defined symmetrical pulse is required.

PULSE DURATION

So far, the emphasis has been placed on the development of a circuit to produce a symmetrical pulse. However, the duration of the pulse is also important. In phase control, the pulse must either trigger a circuit or reset a timing circuit, and the designer will know the pulse duration required. In burst-firing applications with resistive loads, the pulse will normally be used, through a buffer, to trigger the triac or thyristor directly. In such cases, it is generally considered that a pulse duration of between 100 and 200μs will ensure that a triac remains on at the end of one half-cycle, and is held on at the beginning of the next, until the load current exceeds the holding current. This pulse duration thus fulfils the requirement that the trigger pulse is present whenever the load current is not much greater than the holding current.

In all the circuits described, the duration of the output pulse may be adjusted by changing the resistance values suitably.

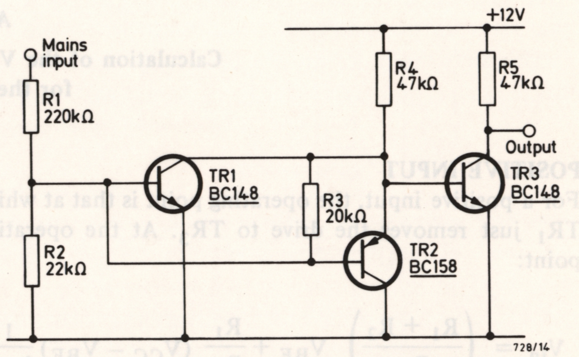

Fig.14 — Final zero-crossing detector circuit

REFERENCE

1. 'Two-point NTC thermistors for measurement and control', Technical Information 1, TP1466/1, Mullard Limited, 1974.

The Appendix is overleaf.

APPENDIX 1

Calculation of the Value of R_3 and the Pulse Width for the Circuit of Fig.14

POSITIVE INPUT

For a positive input, the operating point is that at which TR_1 just removes the drive to TR_3. At the operating point:

$$V_{in} = \left(\frac{R_1 + R_2}{R_2}\right) V_{BE} + \frac{R_1}{R_4} (V_{CC} - V_{BE}) \frac{1}{h_{FE}}, \qquad \ldots(7)$$

where V_{in} is the voltage threshold for a positive input, V_{BE} is the base-emitter voltage of TR_1, V_{CC} is the supply voltage, and h_{FE} the current gain of TR_1.

Assuming that $V_{BE} = 0.7V$ and $h_{FE} = 200$ for the particular transistor considered, and substituting the component values from Fig.14 in Eq.7, the input voltage threshold is given by:

$$V_{in} = \frac{242 \times 10^3}{22 \times 10^3} \times 0.7 + \frac{220 \times 10^3 \times 11.3}{47 \times 10^3 \times 200}, \qquad \ldots(8)$$

$$= 7.96V.$$

The resistor R_3 does not appear in Eq.7 because the base-emitter voltages of TR_1 and TR_3 will be the same at the operating point.

NEGATIVE INPUT

For a negative input, at the operating point the input voltage threshold is given by:

$$V_{in} = \left(\frac{R_1}{R_3}\right) V_{BE} + \frac{R_1}{R_4} (V_{CC} - V_{BE}) \frac{1}{h_{FE}}. \qquad \ldots(9)$$

Assuming that the base-emitter voltages of TR_2 and TR_3 are the same at equal currents, R_2 has no voltage across it at the operating point, and so does not appear in Eq.9.

Equating Eqs.7 and 9 gives:

$$\left(\frac{R_1 + R_2}{R_2}\right) V_{BE} = \frac{R_1}{R_3} \cdot V_{BE}. \qquad \ldots(10)$$

Substituting the component values from Fig.14:

$$R_3 = \frac{220 \times 22 \times 10^3}{242},$$

$$= 20k\Omega \text{ (the value shown for } R_3 \text{ in Fig.14).}$$

PULSE DURATION

The duration of the pulse may also be calculated. The input voltage threshold is given by:

$$V_{in} = V_{pk} \sin \pi f t \qquad \ldots(11)$$

where V_{pk} is the amplitude of the mains supply, f is the mains supply frequency, and t is the pulse duration.

Substituting the appropriate values of V_{in} (7.96V) and the mains amplitude and frequency (340V peak, 50Hz) in Eq.11, the pulse duration is 150μs.

The TDA1050 IC in car radio a.m. receivers

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The TDA1050 car radio IC and its application in two circuits, one with a tuned r.f. stage and the other with an aperiodically-coupled r.f. stage, are described.

INTRODUCTION

Two car radio circuits using the TDA1050 integrated circuit are described. The first incorporates a tuned r.f. stage with two double-tuned circuits for i.f. selectivity. The second also uses two double-tuned circuits for i.f. selectivity but has an aperiodically-coupled r.f. stage. Both circuits feature a wide a.g.c. range, low distortion, and good signal handling, with virtually no change in performance for line voltages between 9 and 16V and ambient temperatures between −25 and +75°C.

This article describes the TDA1050 car radio integrated circuit, discusses its use in the two circuits, and gives full performance data.

DESCRIPTION OF THE TDA1050

The TDA1050 is an integrated circuit for use in high-quality a.m. car radio receivers. It incorporates an r.f. stage, oscillator/mixer, first and second i.f. stages, audio detector, and a separate a.g.c. detector. It also includes an internal voltage stabiliser. The functional block diagram of the IC is shown in Fig.1, together with a

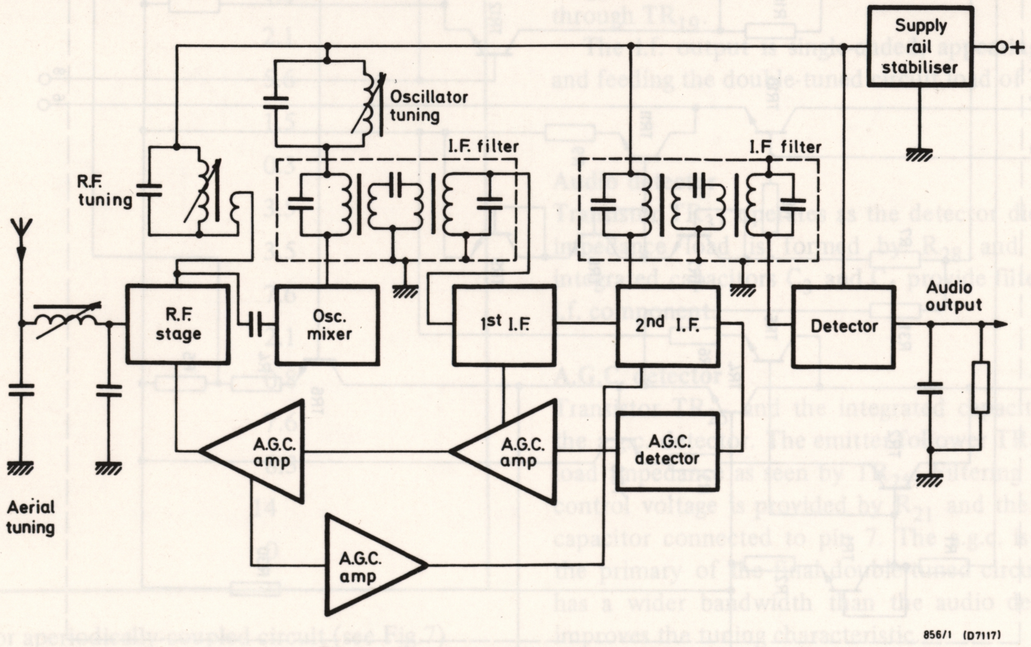


Fig.1 — Circuit functions of TDA1050

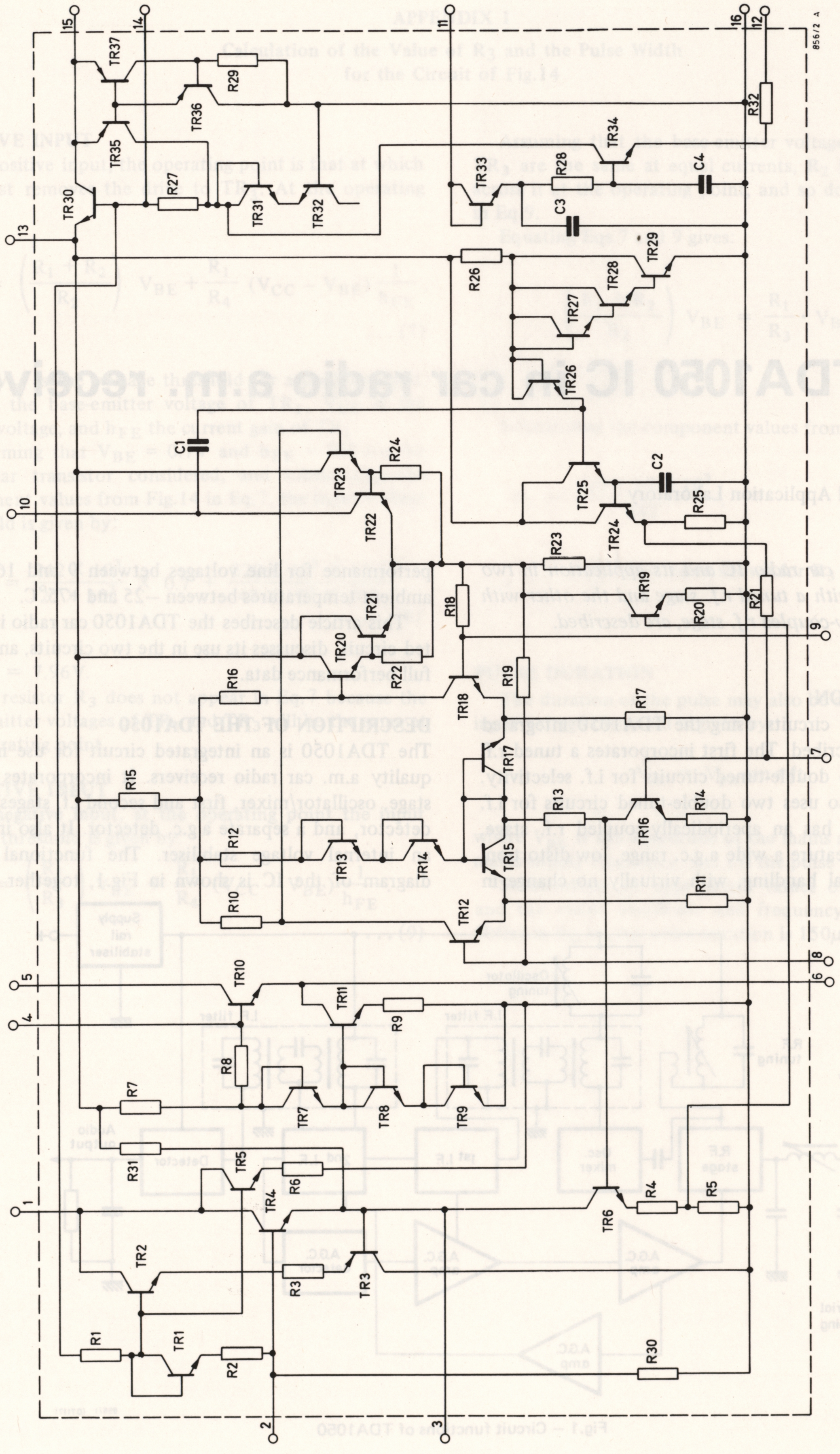


Fig.2 – Circuit diagram of TDA1050

simplified indication of the necessary external circuitry. The circuit diagram of the IC is shown in Fig.2. In the description of the IC that follows, all references to peripheral components are limited to those used in the practical tuned r.f. circuit described later and shown in Fig.3. Pin voltages with respect to earth are given in Table 1.

An internal 7.5V supply rail for the device is provided by an inbuilt stabiliser (see Fig.2). This comprises the reverse-biased emitter-base junction of TR_{32} acting as a voltage regulator diode of about 7.5V, with TR_{31} for V_{be} compensation of the series stabiliser transistor TR_{30}. The stabiliser permits operation of the IC over an input voltage range of 8 to 16V.

R.F. stage

The aerial signal is passed via a π-input filter to the base of the r.f. transistor TR_4 whose emitter current is determined by the a.g.c. transistor TR_6. In a.g.c., TR_4 is reverse-biased, the signal path then being through TR_5. The signal path transition is smoothed by TR_2, R_3, and TR_3. The r.f. stage load is a tuned circuit presenting a nominal 300Ω load at 1MHz. Coupling to the mixer stage is via an inductive tap.

TABLE 1
D.C. Voltages at Pins of TDA1050 with a 14V supply at pin 15

Pin	Approximate d.c. voltage with respect to pin 16 (V)
1	7.6*
2	4.7
3	4.0
4	2.1
5	5.6
6	1.5
7	0.3
8	3.5
9	3.5
10	7.6
11	2.1
12	0.8
13	7.6
14	8.3
15	14
16	0

*6.8V for aperiodically-coupled circuit (see Fig.7)

Mixer/oscillator

Transistor TR_{10} is the self-oscillating mixer transistor whose emitter current is determined by TR_{11} and R_9. The frequency of oscillation is determined by L_6 and its associated components (see Fig.3). A capacitive tap across this tuned circuit feeds the oscillator signal to the emitter of TR_{10}, via a 33Ω resistor at pin 6. The purpose of this resistor is to improve the signal handling at the mixer base; it is left as an external component to increase the flexibility of the circuit. The mixer load is the double-tuned circuit T_1 and T_2, which provides a high impedance to the i.f. frequency and a low impedance to the oscillator frequency.

I.F. stages

The i.f. gain is provided by two d.c.-coupled stages, each provided with a.g.c. Each stage is a balanced long-tail-pair arrangement which has the advantages of removing the need for coupling capacitors to the input and minimising unwanted feedback to the 7.5V line.

Transistors TR_{12} and TR_{18} are the first i.f. stage transistors. Transistors TR_{15} and TR_{17} are normally conducting, and provide a low-impedance path between the emitters of TR_{12} and TR_{18}. A.G.C. is provided by turning on TR_{16}, which extracts current from TR_{15} and TR_{17}, thereby increasing their slope resistances. The second i.f. stage transistors are TR_{20}, TR_{21}, TR_{22}, and TR_{23}, a.g.c. being provided by reducing the current through TR_{19}.

The i.f. output is single-ended, appearing at pin 10, and feeding the double-tuned circuit load of T_3 and T_4.

Audio detector

Transistor TR_{33} operates as the detector diode. A high-impedance load is formed by R_{28} and TR_{34}. The integrated capacitors C_3 and C_4 provide filtering for the i.f. component.

A.G.C. detector

Transistor TR_{25} and the integrated capacitor C_2 form the a.g.c. detector. The emitter-follower TR_{24} raises the load impedance as seen by TR_{25}. Filtering of the a.g.c. control voltage is provided by R_{21} and the electrolytic capacitor connected to pin 7. The a.g.c. is taken from the primary of the final double-tuned circuit so that it has a wider bandwidth than the audio detector. This improves the tuning characteristic.

TUNED R.F. STAGE WITH TWO DOUBLE-TUNED CIRCUITS FOR I.F. SELECTIVITY

The circuit diagram is shown in Fig.3, and a suitable component layout is shown in Fig.4. The tracks from T_3 to pin 10 and from T_4 to pin 11 must each be as short as possible to minimise i.f. feedback. The permeability tuner used with the laboratory prototypes had the following characteristics.

Aerial coil	Q = 75 at 525kHz,	Q = 67 at 1620kHz.
R.F. coil	Q = 66 at 525kHz,	Q = 64 at 1620kHz.
Oscillator coil	Q = 54 at 995kHz,	Q = 37 at 2090kHz.

The r.f. coil tap showed an impedance of 300Ω at 1MHz. The use of tuner units with r.f. coil tap impedances very much greater than that used here is not recommended, as the coil Q-factor may be damped by the mixer input impedance.

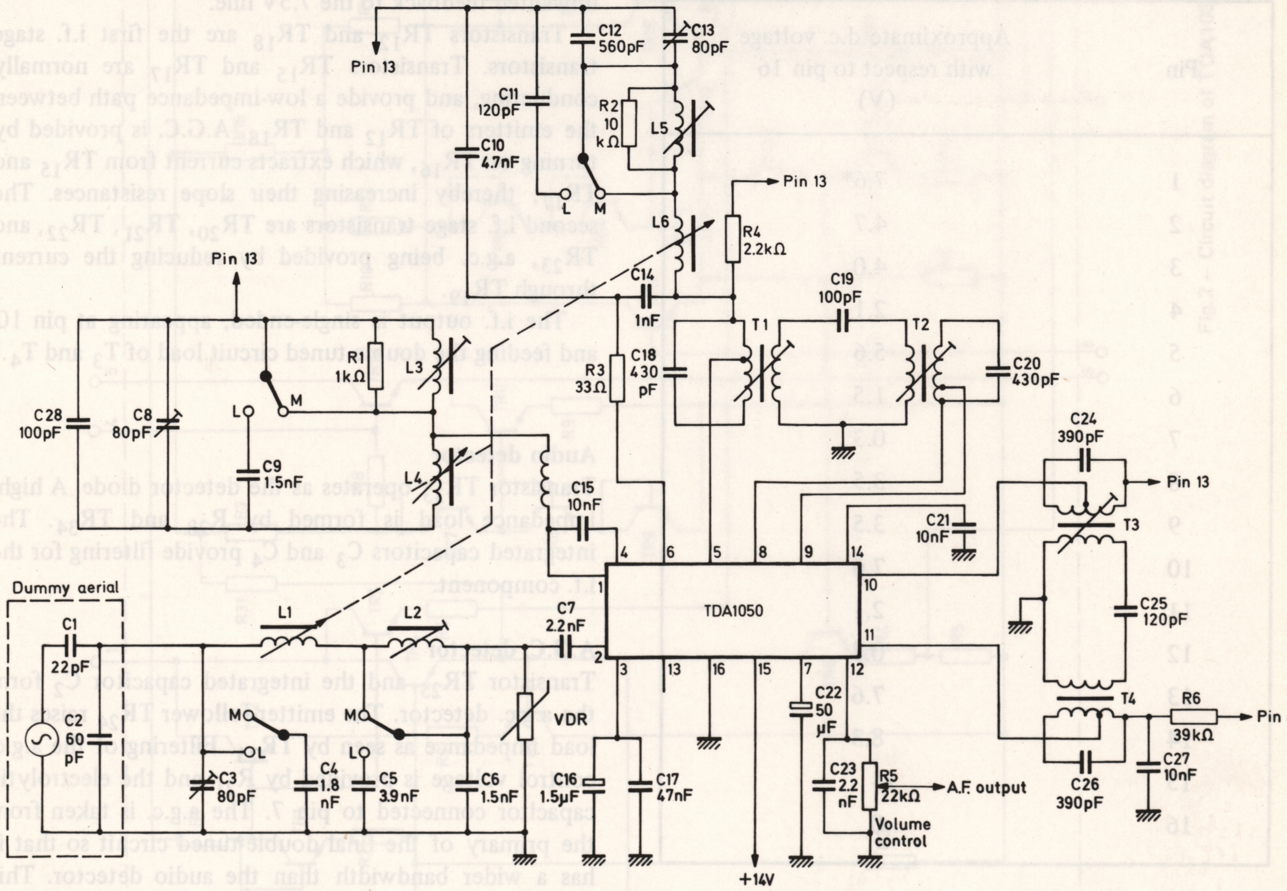
Winding details for the double-tuned i.f. coils and the long-wave coils are given in Appendix 1.

Performance

Plots of audio output, noise, and distortion against r.f. input are shown in Fig.5. The tuning characteristic is shown in Fig.6. Other measured data are given in Table 2.

TABLE 2
Measured Performance of Tuned R.F. Circuit shown in Fig.3

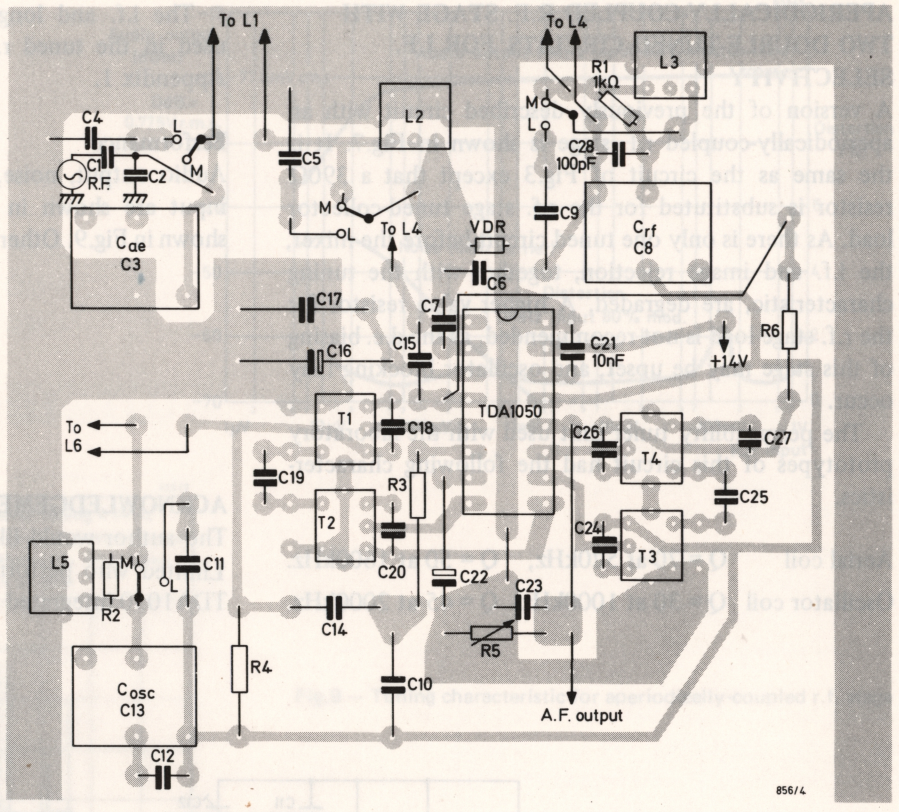
Sensitivity at dummy aerial (20mV audio output for 1MHz input, 80% mod. at 1kHz)	2μV
Audio 3dB bandwidth (1MHz r.f. input at 1mV, 30% mod. ref. 1kHz)	3.3kHz
50dB (S + N)/N ratio (30% mod.)	1mV
I.F. 3dB bandwidth	5kHz
Selectivity ±S9	30dB
Oscillator injection (1MHz receiver tuning frequency)	140mV
I.F. rejection:	
tuned frequency 550kHz	40dB
tuned frequency 1000kHz	65dB
tuned frequency 1500kHz	67dB
Image rejection:	
tuned frequency 550kHz	>97dB
tuned frequency 1000kHz	90dB
tuned frequency 1500kHz	75dB



Pin 13
C12 560pF
C13 80pF
C11 120pF
R2 10 kΩ
L5
C10 4.7nF
L M
Pin 13
L6
R4 2.2kΩ
C14 1nF
C19 100pF
T1
T2
C18 430 pF
R3 33Ω
C20 430pF
Pin 13
R1 1kΩ
L3
C28 100pF
C8 80pF
L M
C9 1.5nF
L4
C15 10nF
C21 10nF
C24 390pF
Pin 13
T3
TDA1050
C25 120pF
Dummy aerial
C1 22pF
60 pF
L1
L2
C7 2.2nF
C3 80pF
C4 1.8 nF
C5 3.9nF
C6 1.5nF
VDR
C16 1.5μF
C17 47nF
C22 50 μF
C23 2.2 nF
R5 22kΩ
A.F. output
Volume control
+14V supply
T4
R6 39kΩ
Pin 4
C26 390pF
C27 10nF
856/3

Fig.3 – Tuned r.f. stage with two double-tuned circuits for i.f. selectivity

Fig.4 – Component layout for tuned r.f. circuit shown in Fig.3



To L1
To L4
C4
C1
R.F.
C2
C5
L2
R1 1kΩ
L3
C28 100pF
C9
To L4
VDR
C6
Cae C3
Crf C8
C17
C7
C16
C15
C21 10nF
R6
+14V
To L6
T1
C18
TDA1050
C26
C27
T4
C19
R3
C25
T2
C20
C24
T3
L5
R2
C11
C22
C23
C14
R5
R4
Cosc C13
C10
A.F. output
C12
856/4

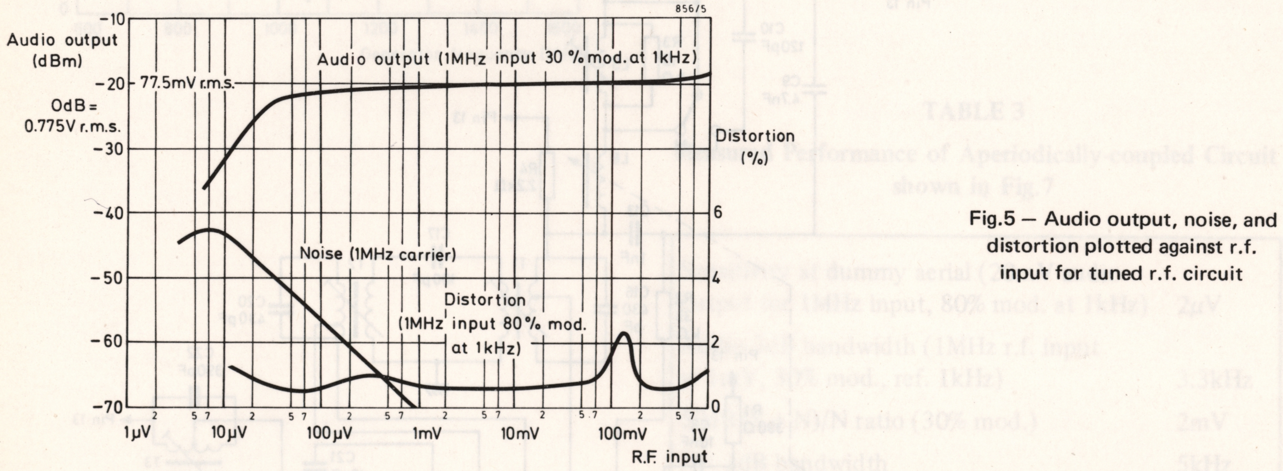


Fig.5 – Audio output, noise, and distortion plotted against r.f. input for tuned r.f. circuit

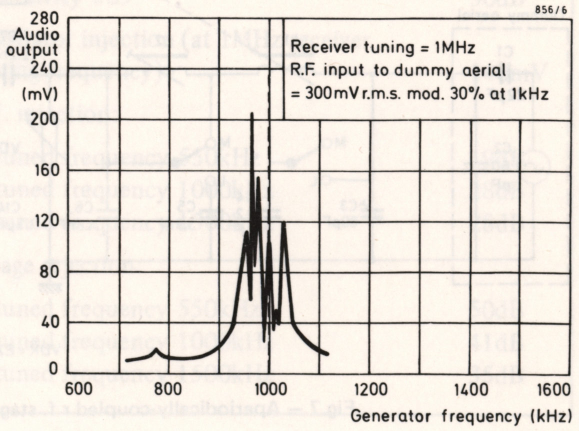


Fig.6 – Tuning characteristic for tuned r.f. circuit

APERIODICALLY-COUPLED R.F. STAGE WITH TWO DOUBLE-TUNED CIRCUITS FOR I.F. SELECTIVITY

A version of the previously described circuit with an aperiodically-coupled r.f. stage is shown in Fig.7. It is the same as the circuit of Fig.3 except that a 390Ω resistor is substituted for the r.f. stage tuned-collector load. As there is only one tuned circuit before the mixer, the i.f. and image rejection, together with the tuning characteristic, are degraded. A higher value resistor for the r.f. stage load is not recommended, as the d.c. biasing of this stage may be upset, and oscillator blocking may occur.

The permeability tuner unit used with the laboratory prototypes of this circuit had the following characteristics.

Aerial coil	Q = 70 at 520kHz,	Q = 50 at 1600kHz.
Oscillator coil	Q = 30 at 1000kHz,	Q = 45 at 2000kHz.

The i.f. and long-wave coils are identical to those used in the tuned r.f. circuit, and details are given in Appendix 1.

Performance

Audio output, noise, and distortion plotted against r.f. input are shown in Fig.8. The tuning characteristic is shown in Fig.9. Other measured data are given in Table 3.

ACKNOWLEDGEMENT

The author would like to thank R. Osborne of Mullard Limited who performed the initial design work on the TDA1050 integrated circuit.

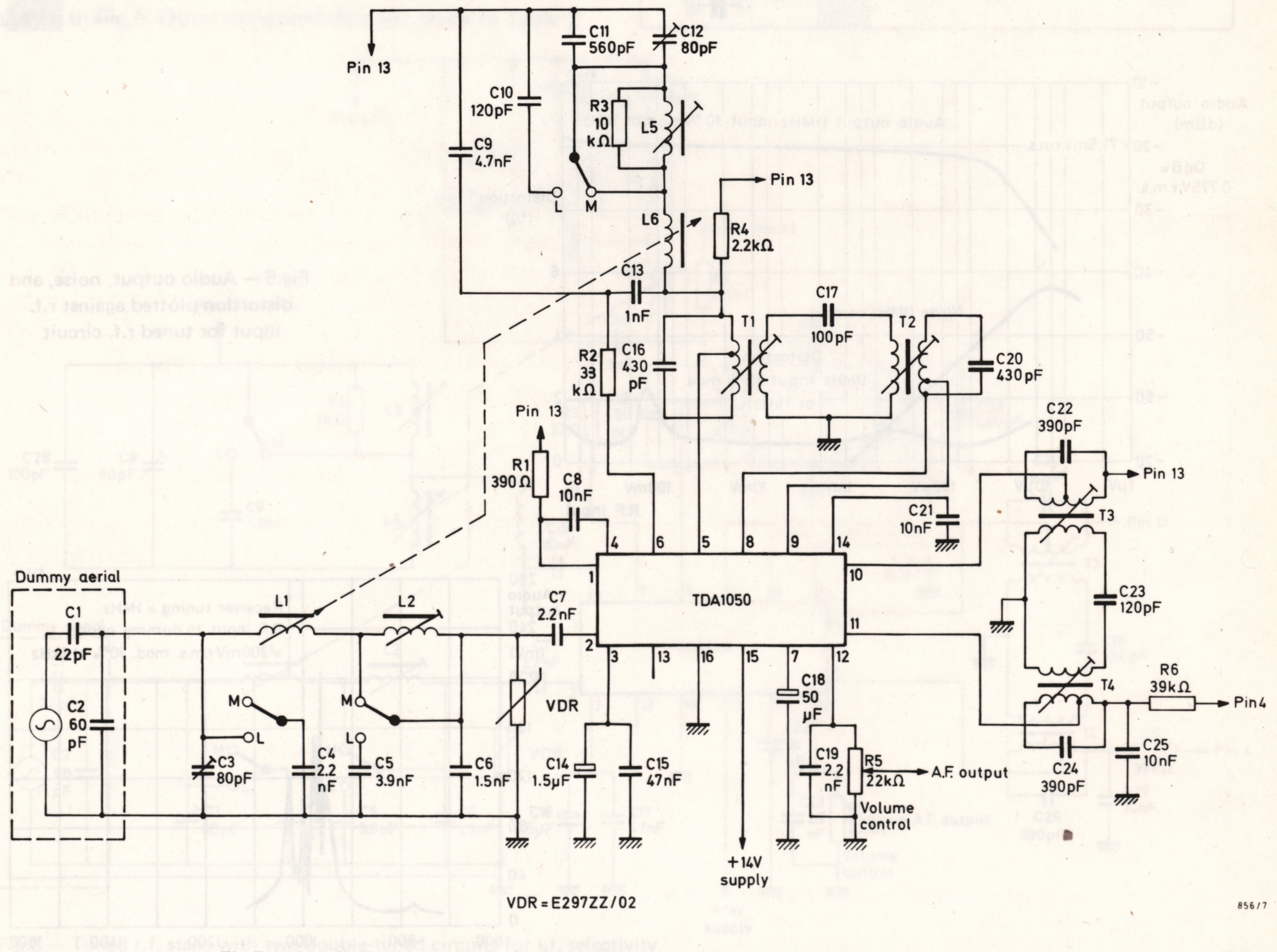
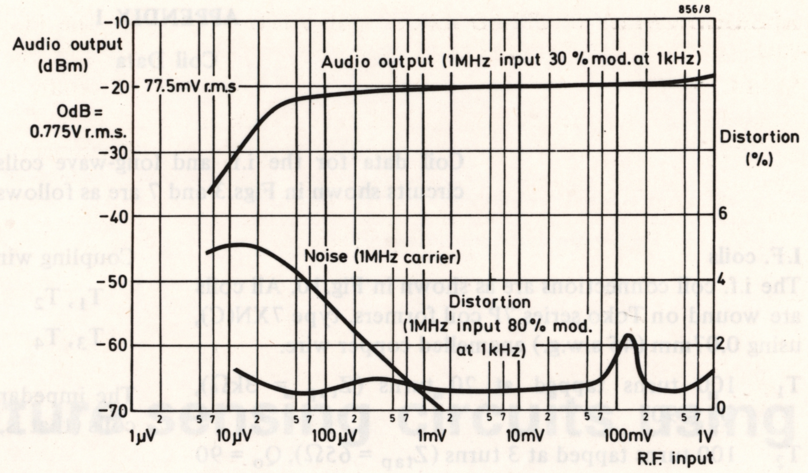

Pin 13
C11 560pF
C12 80pF
C10 120pF
C9 4.7nF
R3 10 kΩ
L5
L M
Pin 13
L6
R4 2.2kΩ
C13 1nF
T1
C17 100pF
T2
R2 33 kΩ
C16 430 pF
C20 430pF
C22 390pF
Pin 13
Pin 13
R1 390Ω
C8 10nF
C21 10nF
T3
4 6 5 8 9 14
1 10
TDA1050
Dummy aerial
C1 22pF
L1
L2
C7 2.2nF
2 11
3 13 16 15 7 12
C23 120pF
C2 60 pF
M O L
M L
C3 80pF
C4 2.2 nF
C5 3.9nF
C6 1.5nF
VDR
C14 1.5µF
C15 47nF
C18 50 µF
C19 2.2 nF
R5 22kΩ
A.F. output
Volume control
T4
R6 39kΩ
Pin4
C24 390pF
C25 10nF
+14V supply
VDR = E297ZZ/02
856/7

Fig.7 — Aperiodically-coupled r.f. stage with two double-tuned circuits for i.f. selectivity

Fig.8 — Audio output, noise, and distortion plotted against r.f. input for aperiodically-coupled r.f. stage



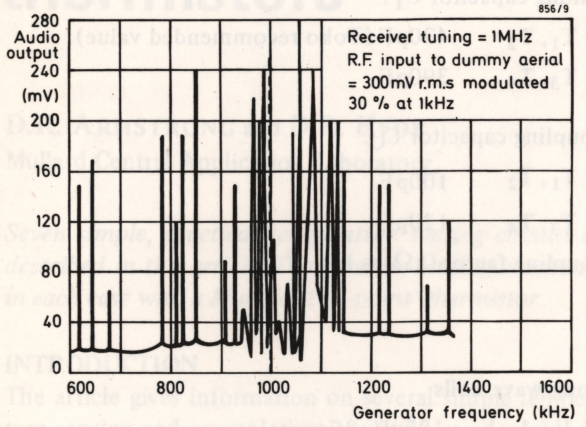


Fig.9 — Tuning characteristic for aperiodically-coupled r.f. stage

TABLE 3

Measured Performance of Aperiodically-coupled Circuit shown in Fig.7

Sensitivity at dummy aerial (20mV audio output for 1MHz input, 80% mod. at 1kHz)	2µV
Audio 3dB bandwidth (1MHz r.f. input at 1mV, 30% mod., ref. 1kHz)	3.3kHz
50dB (S + N)/N ratio (30% mod.)	2mV
I.F. 3dB bandwidth	5kHz
Selectivity ±S9	30dB
Oscillator injection (at 1MHz receiver tuning frequency)	140mV
I.F. rejection:	
tuned frequency 550kHz	19dB
tuned frequency 1000kHz	28dB
tuned frequency 1500kHz	28dB
Image rejection:	
tuned frequency 550kHz	50dB
tuned frequency 1000kHz	41dB
tuned frequency 1500kHz	35dB

APPENDIX 1

Coil Data

Coil data for the i.f. and long-wave coils used in the circuits shown in Figs.3 and 7 are as follows.

I.F. coils

The i.f. coil connections are as shown in Fig.10. All coils are wound on Toko series 7P coil formers, type 7XN(C), using 0.07mm (45 s.w.g.) enamelled copper wire.

T_1	100 turns tapped at 20 turns (Z_{tap} = 3kΩ), Q_o = 90
T_2	100 turns tapped at 3 turns (Z_{tap} = 65Ω), Q_o = 90
T_3	100 turns tapped at 75 turns (Z_{tap} = 57kΩ), Q_o = 100
T_4	100 turns tapped at 75 turns (Z_{tap} = 57kΩ), Q_o = 100

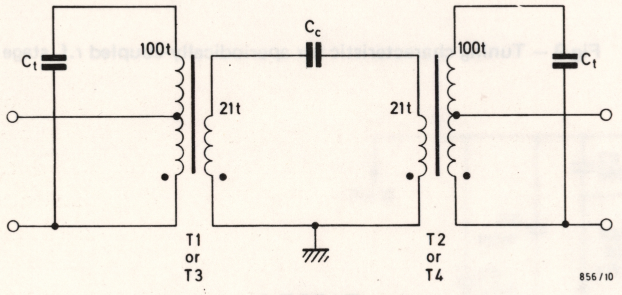


Fig.10 — I.F. coil connections

Coupling winding impedances:

T_1, T_2	3.2kΩ
T_3, T_4	4.2kΩ

The impedances and Q values given above are for single coils (that is, not coupled).

Tuning capacitor C_T:

T_1, T_2	430pF (Toko recommended value)
T_3, T_4	390pF

Coupling capacitor C_C:

T_1, T_2	100pF
T_3, T_4	120pF

Coupling factor (kΩ) = 1

Long-wave coils

L_2, L_3	185μH, 85 turns
L_5	127μH, 71 turns

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Vol.14, No.131, July 1976

Microwave doppler intruder alarms

Page 51: In the right-hand column, 8 lines from the bottom of the page, 1.5dB should read 11.5dB.

Simple temperature sensing circuits using thermistors

D.R. Armstrong and D.R. Hyde
Mullard Central Application Laboratory

Seven simple, practical temperature sensing circuits are described in this article. Temperature sensing is achieved in each case with a Mullard 'two-point' thermistor.

INTRODUCTION

The article gives information on several simple temperature sensing and control circuits which are adaptable for a number of applications, including the circuit of a thermostat specifically intended for the control of wash temperature in automatic washing machines. The seven circuits described are as follows:

1) simple thermometer,
2) simple bridge thermometer,
3) differential thermometer,
4) ice-point warning device,
5) general-purpose mains-isolated thermostat,
6) circuit for automatic switch-off of heater load,
7) thermostat for automatic washing machines.

The circuits all use devices from the range of plastic-encapsulated 'two-point' n.t.c. thermistors which are particularly suitable for temperature sensing applications owing to their accuracy; the tolerance on resistance does not exceed ±8% of the nominal value in any instance (Ref.1). Furthermore, their sensitivity allows their use with a minimum of amplification.

The circuits shown assume nominal values throughout, and therefore tolerances should be checked by the user. The thermistor is selected according to the temperature range of each circuit, as shown in Table 1.

SIMPLE THERMOMETER

A simple thermometer circuit using a two-point thermistor is shown in Fig.1. The range of the thermometer is from −20 to +24°C. Because the thermistor resistance approaches an inverse exponential characteristic, the thermometer is sensibly linear over the range 0 to +20°C. Meter scale calibration information which relates to the

TABLE 1

Brief Specification of 'Two-point' Thermistor Range

Type No. 2322....	Colour code	Temperature range (°C)	Resistance (kΩ)	at temperature (°C)	Resistance (kΩ)	at temperature (°C)
640 90002	blue	−20 to +25	8.6 ± 8%	−20	1.215 ± 7%	+25
640 90003	white	−30 to −10	13.35 ± 5%	−30	5 ± 4.5%	−10
640 90004	grey	+25 to +100	12 ± 7%	+25	0.95 ± 5%	+100
640 90005	black	+100 to +200	16.7 ± 7%	+100	1.12 ± 7%	+200

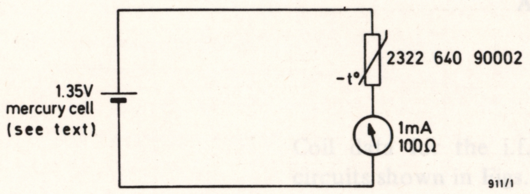

Fig.1 — Simple thermometer circuit

nominal resistance of the 2322 640 90002 thermistor is shown in Fig.2 and Table 2. This calibration information applies to a 1mA f.s.d. moving-coil meter with an internal resistance of 100Ω. An adequately stable reference voltage is provided for the thermometer by a 1.35V mercury cell. In the calculations for Fig.2 and Table 2, the cell was assumed to have negligible internal resistance. Alterations to these circuit conditions must be compensated in the calibration of the meter scale. Alternative temperature ranges can be provided by use of the other two-point thermistors.

TABLE 2

Values of Thermistor Resistance and Meter Current over a Range of Temperatures

Temperature (°C)	2322 640 90002 nominal thermistor resistance (Ω)	Meter current (mA)
−20	8600	0.155
−15	6690	0.199
−10	5254	0.252
−5	4163	0.317
0	3327	0.394
+5	2681	0.485
+10	2176	0.593
+15	1780	0.718
+20	1465	0.863
+24	1260	0.993

SIMPLE BRIDGE (FREEZER) THERMOMETER

A better circuit for a thermometer is shown in Fig.3. The component values are chosen to give balance at a temperature of −8°C, and the meter monitors imbalance of the bridge to indicate temperatures in the range −18 to 0°C. A meter calibration curve is shown in Fig.4, and this is plotted from values given in Table 3.

The bridge is accurate and independent of the supply voltage when it is in the balanced condition at approximately −8°C, but will be dependent on the battery voltage at other temperatures. The use of three mercury cells for the supply is therefore suggested in this application.

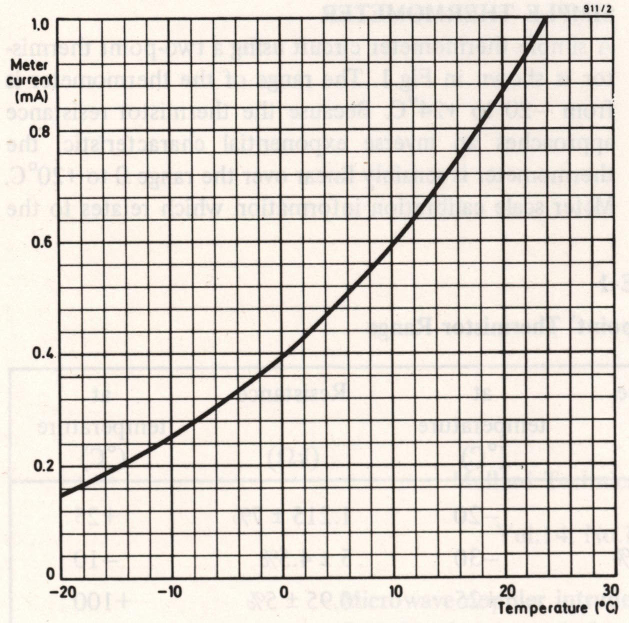

Fig.2 — Calibration of meter scale, from mA to °C, for simple thermometer

DIFFERENTIAL TEMPERATURE INDICATOR

Fig.5 shows a simple differential temperature indicator circuit which provides a useful indication of thermal stability in temperature-controlled environments (for example, laboratory ovens).

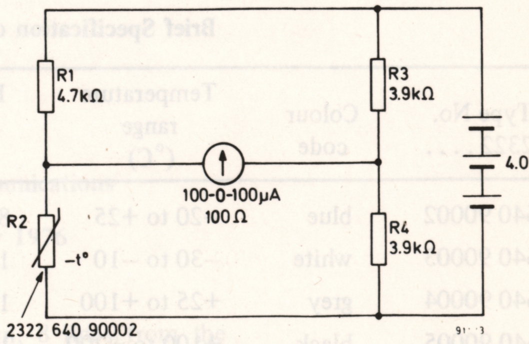

Fig.3 — Simple bridge (freezer) thermometer circuit

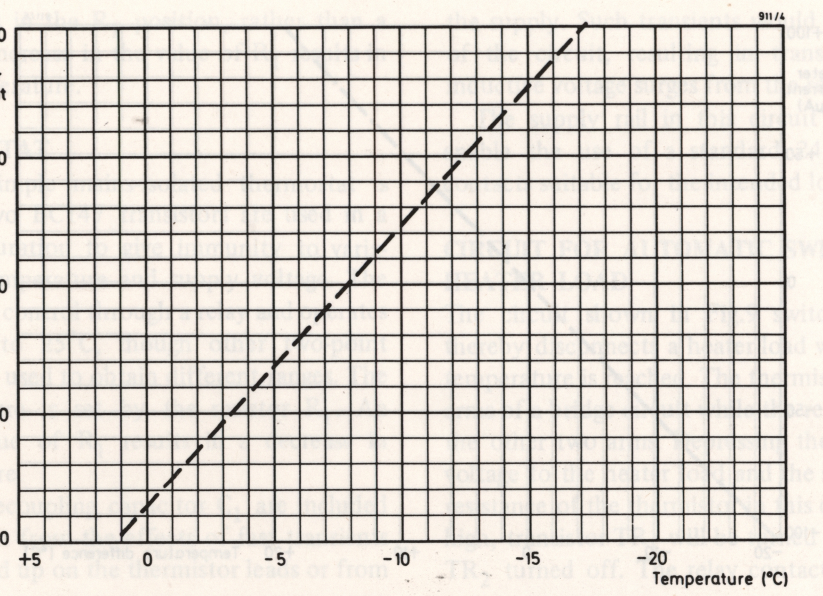

Fig.4 — Calibration of meter scale, from µA to °C, for bridge thermometer

TABLE 3

Values of Thermistor Resistance and Meter Current over a Range of Temperatures

Temperature (°C)	2322 640 90002 nominal thermistor resistance (Ω)	Meter current (µA)
+1	3184	−99
0	3327	−87
−2	3636	−63
−4	3978	−40
−6	4358	−18
−7	4564	−7
−8	4782	4
−10	5253	25
−12	5780	45
−14	6369	64
−16	7029	83
−18	7768	100
−20	8600	117

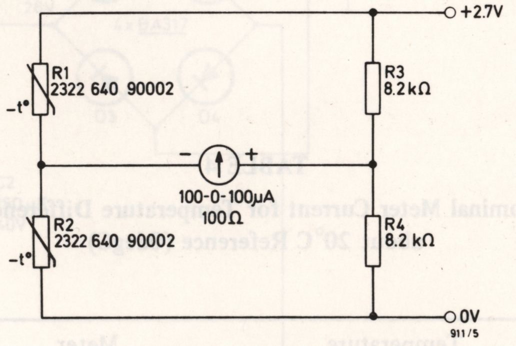

Fig.5 — Differential temperature indicator circuit

A meter scale calibration curve is shown in Fig.6, plotted from the values given in Table 4. These values relate to nominal 2322 640 90002 thermistors, where thermistor R_1 is at a reference temperature of 20°C, and thermistor R_2 is subjected to temperature changes above and below 20°C. The calibration accuracy depends on the operating temperatures and the error introduced by the meter current. The circuit gives an approximately linear characteristic, but error is introduced by the varying parallel resistance of the two thermistors. In this circuit, the error is about 5% at f.s.d. The values shown in Table 4 were calculated for a 100-0-100µA f.s.d. moving-coil meter with an internal resistance of 100Ω, and a power supply (two mercury cells) having negligible resistance. Alterations to these circuit conditions must be compensated in the calibration of the meter scale.

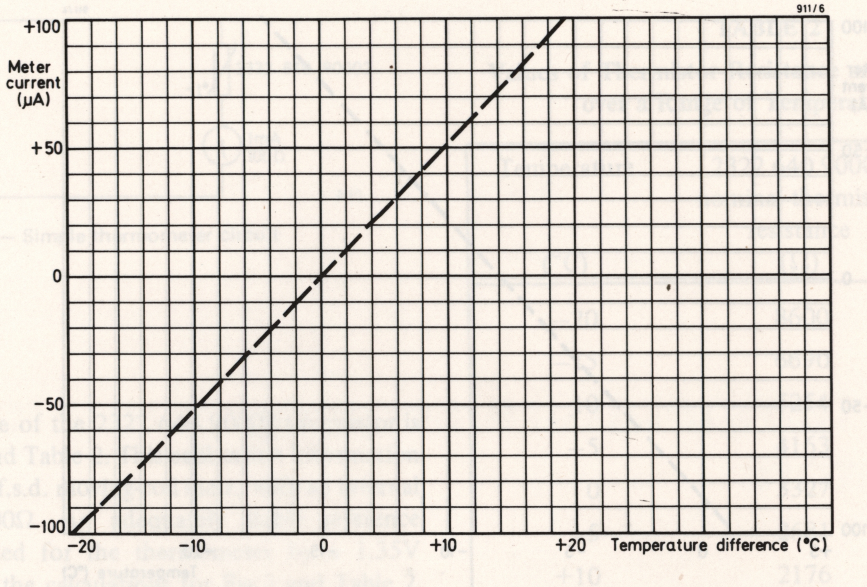

Fig.6 – Calibration of meter scale, from µA to °C, for differential temperature indicator

TABLE 4

Nominal Meter Current for Temperature Differences about 20°C Reference (0degC)

Temperature difference (degC)	Meter current (µA)
−20	−101
−18	−92
−16	−82
−14	−73
−12	−63
−10	−52
−8	−42
−6	−32
−4	−21
−2	−11
0	0
+2	+11
+4	+21
+5	+26

ICE-POINT WARNING DEVICE

The circuit of an ice-point warning device is shown in Fig.7. A visible warning of freezing conditions is given by the device. This circuit is intended for use in the car or in other applications requiring a temperature warning device, and the circuit can be adjusted to operate at temperatures above or below freezing point.

A long-tail pair is used in the circuit to sense resistance changes of thermistor R_3 (2322 640 90002) which forms one arm of a bridge. Some hysteresis is applied by resistor R_8, to prevent excessive dissipation in the transistor TR_3. The circuit switches the lamp on when the temperature falls to 0°C. If required, the operating point can be adjusted by changing the value of resistor R_1. This resistor is included in series with R_2 so that small changes in sensing temperature may be made, using fixed resistors in the R_1 position, rather than a potentiometer. An increase in the value of R_1 results in a lower sensing temperature.

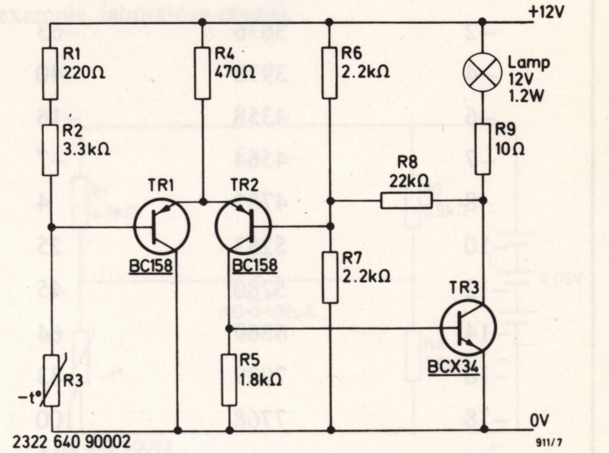
+12V
R1
220Ω
R2
3.3kΩ
R3
−t°
2322 640 90002
R4
470Ω
TR1
BC158
TR2
BC158
R5
1.8kΩ
R6
2.2kΩ
R7
2.2kΩ
R8
22kΩ
Lamp
12V
1.2W
R9
10Ω
TR3
BCX34
0V
911/7

Fig.7 – Ice-point warning device

SIMPLE THERMOSTAT

The circuit of a simple mains-isolated thermostat is shown in Fig.8. Two BC147 transistors are used in a long-tail-pair configuration to give immunity to variations in ambient temperature and supply voltage. The thermostat exercises control through a relay and operates over the range 0 to 25°C, though other two-point thermistors could be used to obtain different ranges. The operating temperature is set by the resistor R_1. An increase in the value of R_1 results in a decrease in operating temperature.

Diode D_1 and decoupling capacitor C_1 are included to protect the circuit from the effects of fast transients which may be picked up on the thermistor leads or from the supply. Such transients would cause rapid switching of the circuit, resulting in transistor damage due to inductive voltage surges from the relay.

The supply rail in this circuit has been chosen to enable the use of a standard 24V 580Ω relay, with contacts suitable for the intended load.

CIRCUIT FOR AUTOMATIC SWITCH-OFF OF HEATER LOAD

The circuit shown in Fig.9 switches off a relay and thereby disconnects a heater load when a predetermined temperature is reached. The thermistor and R_1 form two arms of a bridge circuit while the relay and R_4 constitute the other two arms. Depressing the start button applies voltage to the heater load and the switching circuit. The resistance of the thermistor in this cold condition will be high, transistor TR_1 will be turned on, and consequently TR_2 turned off. The relay contacts will close, allowing

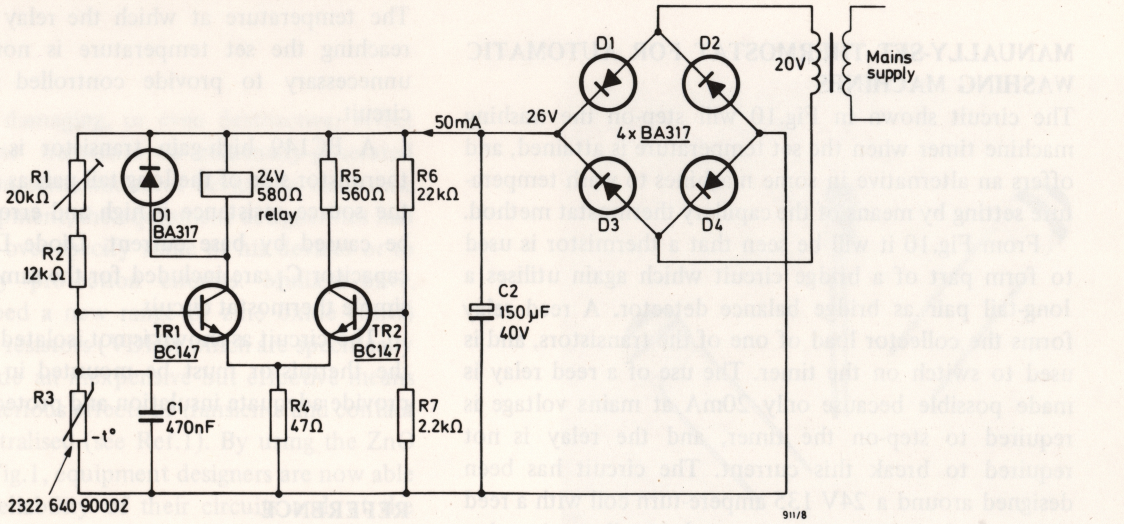
D1
D2
20V
Mains supply
4 x BA317
D3
D4
26V
50mA
R1
20kΩ
D1
BA317
24V
580Ω
relay
R5
560Ω
R6
22kΩ
R2
12kΩ
TR1
BC147
TR2
BC147
C2
150µF
40V
R3
−t°
2322 640 90002
C1
470nF
R4
47Ω
R7
2.2kΩ
911/8

Fig.8 – Simple thermostat, with a mains isolating transformer

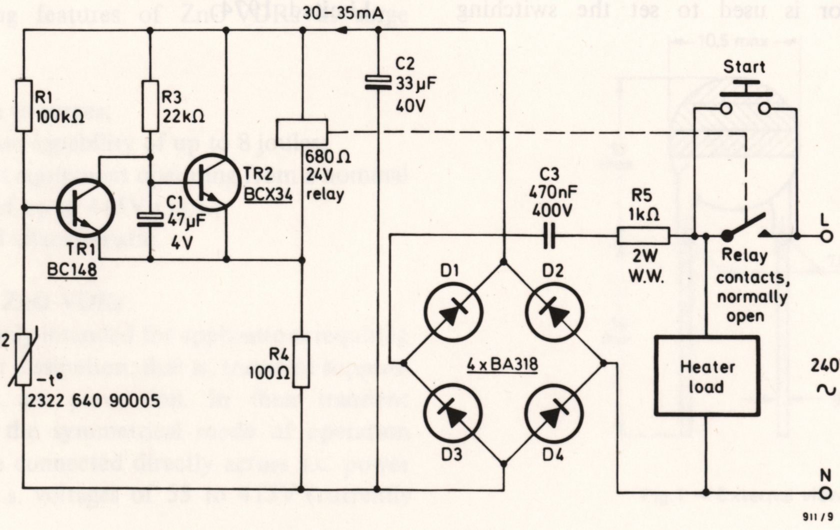
30–35mA
R1
100kΩ
R3
22kΩ
C2
33µF
40V
Start
TR1
BC148
C1
47µF
4V
TR2
BCX34
680Ω
24V
relay
C3
470nF
400V
R5
1kΩ
2W
W.W.
L
Relay contacts, normally open
D1
D2
4 x BA318
D3
D4
R2
−t°
2322 640 90005
R4
100Ω
Heater load
240V
∼
N
911/9

Fig.9 – Circuit for automatic switch-off of heater load

the start button to be released and the heater load to remain switched on.

At the predetermined temperature (approximately 100°C with the value of R_1 and thermistor type shown in this circuit), the resistance of the thermistor decreases to a point at which TR_1 is switched off and TR_2 switched on. The latter then effectively shorts-out the relay and provides positive feedback, since the relay coil is one arm of the bridge. This causes the circuit to regenerate, and the relay drops out quickly and disconnects the supply.

The capacitive dropper, C_3, is used to minimise power dissipation in the circuit, the 1kΩ resistor R_5 being included to limit the current surges in the capacitor and diodes. Capacitor C_1 at the base of TR_2 enables the circuit to be reset when the temperature is close to the switch-off temperature and also prevents the circuit switching off during short transients of interference.

MANUALLY-SET THERMOSTAT FOR AUTOMATIC WASHING MACHINES

The circuit shown in Fig.10 will step-on the washing machine timer when the set temperature is attained, and offers an alternative in some machines to wash temperature setting by means of the capillary thermostat method.

From Fig.10 it will be seen that a thermistor is used to form part of a bridge circuit which again utilises a long-tail pair as bridge balance detector. A reed relay forms the collector load of one of the transistors, and is used to switch on the timer. The use of a reed relay is made possible because only 20mA at mains voltage is required to step-on the timer, and the relay is not required to break this current. The circuit has been designed around a 24V 135 ampere-turn coil with a reed insert of 40 to 60 ampere-turns and this allows the relay to be operated at 15V.

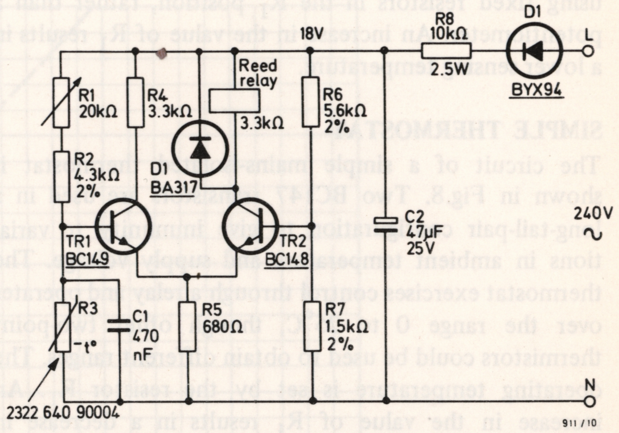

Fig.10 – Washing machine thermostat

A variable resistor is used to set the switching temperature as required within the range 40 to 90°C. The temperature at which the relay opens again after reaching the set temperature is not critical so it is unnecessary to provide controlled hysteresis in the circuit.

A BC149 high-gain transistor is employed on the thermistor side of the long-tail pair as at low temperature the source resistance is high and error would otherwise be caused by base current. Diode D_1 and decoupling capacitor C_1 are included for the same purpose as in the simple thermostat circuit.

The circuit as shown is not isolated from the mains so the thermistor must be mounted in such a way as to provide adequate insulation and protection for the user.

REFERENCE

1. 'Two-point NTC Thermistors for measurement and control', Technical Information 1, TP1466/1, Mullard Limited, 1974.

Product Information: Zinc oxide VDRs for transient suppression

INTRODUCTION

To minimise the damaging, or even destructive, effect which mains-borne transients or internally-generated spikes can have on semiconductor devices in electronic and electrical equipment, the equipment designer has had hitherto either to over-specify some of his devices or to incorporate costly protection circuits. Mullard have, therefore, developed a new range of zinc oxide (ZnO) voltage dependent resistors (VDRs) which are specifically designed to provide an inexpensive but effective means by which the deleterious effects of transients and contact arcing can be neutralised (see Ref.1). By using the ZnO VDRs shown in Fig.1, equipment designers are now able to increase the reliability of their circuits and, at the same time, reduce component costs by specifying semiconductor devices with more moderate ratings.

The outstanding features of ZnO VDRs as surge suppressors are:

- instant response to surges;
- energy absorption capability of up to 8 joules;
- ability to protect equipment operating from a nominal supply voltage of up to 415V r.m.s.;
- symmetrical V/I characteristic.

THE RANGE OF ZnO VDRs

ZnO VDRs are mainly intended for applications requiring intermittent power dissipation; that is, transient suppression and contact arc prevention. In their transient suppression role, the symmetrical mode of operation allows them to be connected directly across a.c. power lines carrying r.m.s. voltages of 55 to 415V (currently

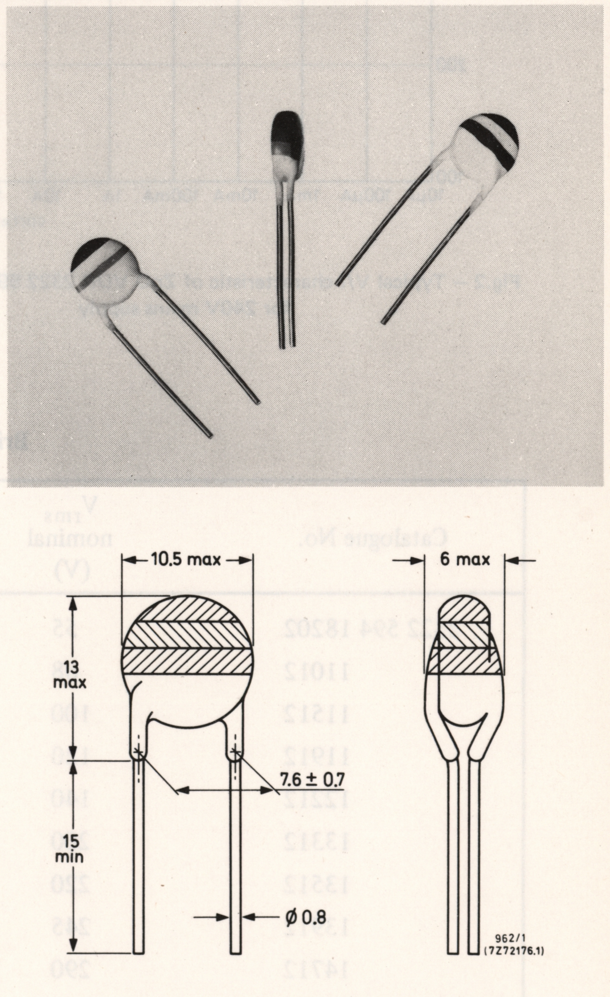

Fig.1 – External view of Mullard ZnO VDRs

available types). They are capable of withstanding voltage or current pulses with a high peak energy level. Brief data for the range is given in Table 1.

Fig.2 shows a typical voltage/current characteristic of one device in the range. This curve also shows the possible variation in the characteristic caused by manufacturing tolerances. The upward turn of the characteristic (decreasing non-linearity) is due to the increasing influence of the linear series resistance as the non-linear resistance falls to very low values at high currents.

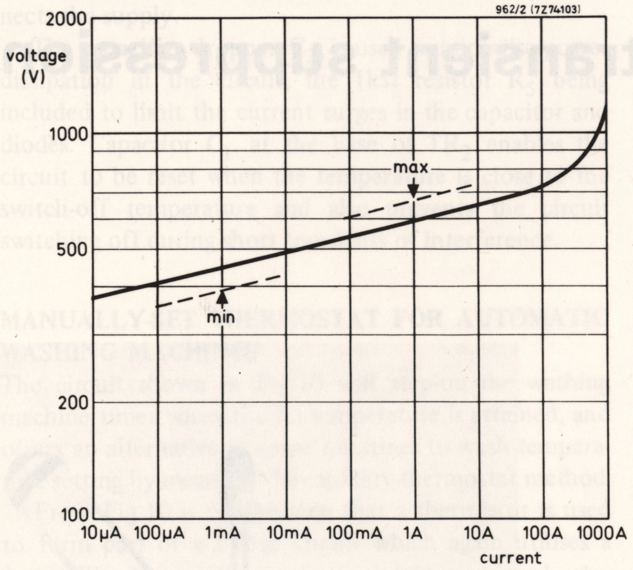

Fig.2 – Typical V/I characteristic of ZnO VDR 2322 594 13912 for 240V mains supply

On the straight portion of the characteristic, the relationship between voltage and current is given by:

$$V = CI^{\beta},$$

where β is the non-linearity index, and C is the voltage across the VDR when it is passing a current of 1A. A typical value of β for a ZnO VDR is 0.035. This means that if the current through the VDR increases by a factor of 10 within the straight-line portion of the curve, the voltage across it increases by a factor of 1.08.

A good approximation of the relationship between current and voltage on the curved portion of the curve is:

$$V = CI^{\beta} + IR_s$$

where R_s is the series resistance of the VDR. Typical values of series resistance range from 0.3Ω for the lower-voltage types up to 1.0Ω for the higher-voltage end of the range. The value is approximately 0.6Ω for types intended for 220V or 240V r.m.s. a.c. supplies.

Further information on ZnO VDRs together with full information on their operation and their application in practical circuits are given in Ref.1. This publication and the full published data on ZnO VDRs are available on request.

REFERENCE

1. 'Zinc oxide voltage dependent resistors for transient suppression', Technical Information 26, TP1543, Mullard Limited, 1976.

TABLE 1

Brief Data for ZnO VDRs

Catalogue No.	V_{rms} nominal (V)	Minimum voltage at 1mA peak (V)	Maximum voltage at 1A peak, C_{max} (V)
2322 594 18202	55	82	140
11012	68	100	170
11512	100	150	255
11912	120	190	325
12212	140	220	375
13312	200	330	565
13512	220	350	600
13912	245	390	665
14712	290	470	805
16212	380	620	1050
16812	415	680	1160